RIDER

TELEVISION MANUAL

HOW IT WORKS

VOLUME 1

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TELEVISION
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PREFACE

IT is our feeling that while television embraces many new techniques, nothing is so complicated in television that makes it beyond the capabilities and comprehension of the members of the service industry.

In line with this thought, we give in this "How It Works" explanations of the operation of television receivers such as are on today's market. A man must know more than the contents of this book, which is merely a stepping-stone, for no textbook can be hoped to convey everything at once. This book will develop a familiarity with television — practice will develop more — other texts will add still more knowledge — attendance at a school will be of great benefit — past experience on regular broadcast receivers is a cornerstone on which to build — in short, it is a combination of many things that will make a competent television technician.

One thing is vital: while an individual may acknowledge that he does not know sufficient at the moment to service a television receiver, he should not develop a fear of what is in such a receiver. It is hoped that the contents of this book will tend to expose what is in a television set so as to show that making the effort to learn will develop the necessary capability.

Foremost in a man's mind should be the fact that unlike a radio receiver which does not have a visual indicator, the picture tube in a television receiver is a guide post toward the very many defects which may be found in a receiver and will tend to lead the man having the necessary power of interpretation, toward the cause of the faults.

We have incorporated in this book the theoretical principles upon which the various circuits of a television receiver are based, for only if a man has a grasp of this underlying theory can he hope to become a competent technician. As theory and practice must go hand-in-hand, we have used as typical examples those circuits or components that are in actual receivers, not forgetting those which might be called unusual. Thus if the reader will digest the contents of these pages, he will find himself with a widened viewpoint that will be valuable in his advancement toward becoming a better all-around television technician.

The opening chapter presents an over-all picture of television, for it was felt the reader should at least have a nodding acquaintance with the manner in which the subject is televised, how the image is built up in the camera tube, how the video and sound signals are transmitted, and then picked up, separated and reproduced in the receiver. The next chapter deals with the television channels and the characteristics of the video signal; operating bandwidth characteristics, and the carrier and intermediate frequencies. The third chapter discusses antennas designed for the reception of television signals, how they should be oriented and how they should be installed. The next eight chapters cover the receiver proper, the titles being the r-f amplifier, oscillator, and converter; the sound channel; the video i-f system and detector; video amplifier and d-c restorer; sync circuits; sweep circuits; the picture tube, and finally the power supplies. The final chapter contains practical discussions of test instruments, signal generators, etc., used in the alignment and maintenance of receivers, alignment procedures, and trouble-shooting suggestions.

We wish to express our gratitude to the many manufacturers who cooperated with us in supplying technical data and descriptions of their products. Also we wish to thank the members of the staff of John F. Rider Publisher who assisted in the preparation and writing of this book and whose names appear at the heads of the chapters.

JOHN F. RIDER

CHAPTER 1

GENERAL ASPECTS OF THE TELEVISION SYSTEM

By William Bouie

No comments can be made on recent advances in the radio art without recognizing that after many years of development, television has finally reached the stage where servicemen can no longer afford to ignore its existence. In most parts of the country television is already an accomplished fact, thousands of receivers have been sold and installed, and all signs point toward the gradual expansion of television facilities into sections which do not have coverage at the present time.

Recognizing, then, that the radio serviceman's job is being extended to include not only radio servicing as we know it today, but also the servicing of television receivers, this section explaining the fundamentals of television is being offered in the belief that it will be of help to many of you in dealing with the problems introduced by television. Throughout you will find that it is written from the viewpoint of the man who is called upon to install and service television receivers; and yet, although there is this emphasis on the practical aspects of television, you will find that an essential amount of theory is included. Lest any of you feel that this inclusion of what might be called "theoretical" material is unnecessary, it should be pointed out that television is an extremely complicated subject, much more so than is radio, and that television receivers cannot be serviced efficiently by any one who does not understand the principles underlying the operation of the system. The different situations which can arise are so numerous that it is impossible to list them, to explain the reasons for each, and in the case of faulty operation to locate the source of the trouble. As in radio servicing, so it is in television servicing: a thorough understanding of the fundamental principles of operation is invaluable.

Fortunately for the radio serviceman, the advent of television does not mean that an entirely new field must be learned. The more you study television, the more you will come to realize that television embodies every principle that has ever been used in radio—and more besides. You will marvel at the ingenuity of television engineers in using the time-proved principles of radio, in adapting these to the needs of television, and in discovering new principles and new techniques wherever the available ones were inadequate. Thus, for example, you will find the simple diode rectifier being used in

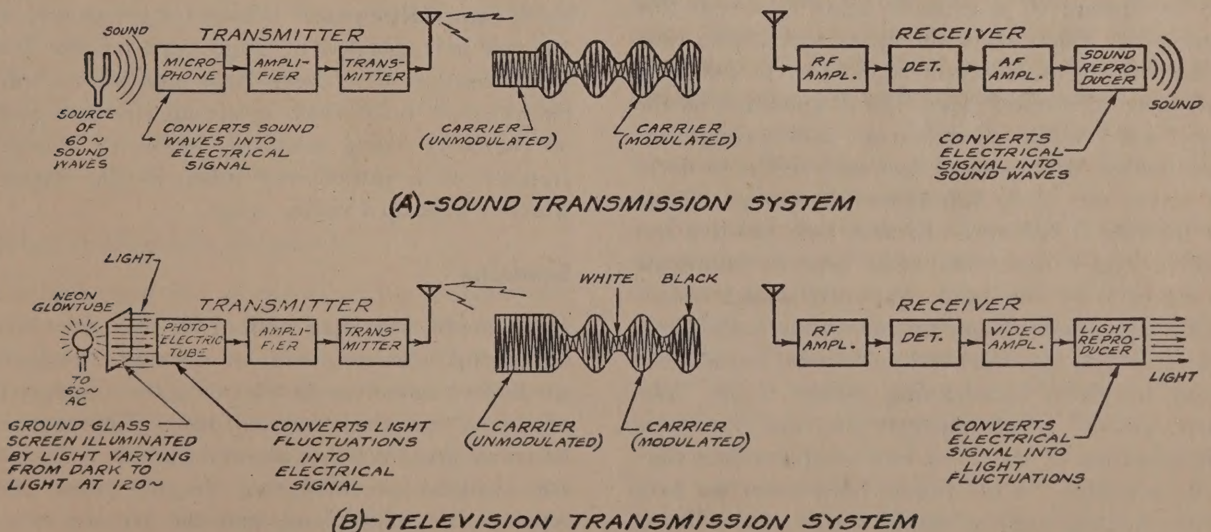

SOUND
TRANSMITTER
MICRO-PHONE
AMPLI-FIER
TRANS-MITTER
SOURCE OF 60 ~ SOUND WAVES
CONVERTS SOUND WAVES INTO ELECTRICAL SIGNAL
CARRIER (UNMODULATED)
CARRIER (MODULATED)
RECEIVER
RF AMPL.
DET.
AF AMPL.
SOUND REPRO-DUCER
SOUND
CONVERTS ELECTRICAL SIGNAL INTO SOUND WAVES
(A)-SOUND TRANSMISSION SYSTEM
LIGHT
NEON GLOWTUBE
TRANSMITTER
PHOTO-ELECTRIC TUBE
AMPLI-FIER
TRANS-MITTER
TO 60~ AC
GROUND GLASS SCREEN ILLUMINATED BY LIGHT VARYING FROM DARK TO LIGHT AT 120~
CONVERTS LIGHT FLUCTUATIONS INTO ELECTRICAL SIGNAL
WHITE
BLACK
CARRIER (UNMODULATED)
CARRIER (MODULATED)
RECEIVER
RF AMPL.
DET.
VIDEO AMPL.
LIGHT REPRO-DUCER
LIGHT
CONVERTS ELECTRICAL SIGNAL INTO LIGHT FLUCTUATIONS
(B)-TELEVISION TRANSMISSION SYSTEM

Fig. 1-1.—Part (A) shows an a-m sound transmission system and (B) shows a television transmission system. In the television system the photoelectric tube replaces the microphone and the light reproducer replaces the speaker.

clipping circuits, in d-c restoring circuits, in video detector circuits, and in many other applications. If you understand how the diode rectifier operates, then you will recognize in each one of these "new" circuits an old friend. True, it will take some time before you become familiar with the need for these circuits and the specific modifications to achieve certain results, still you will be amply repaid for the effort in the new opportunities which television offers to the trained serviceman and technician.

Comparison with Sound Broadcasting

Just as it is the problem of radio broadcasting to recreate sound at places distant from the actual sound, so it is the problem of television to recreate a scene at places distant from the original scene. In the case of amplitude modulated (a-m) sound broadcasting, as Fig. 1-1 (A) shows, sound vibrations produced by, let us say, a tuning fork, are picked up by a *microphone* which converts these vibrations into corresponding electrical vibrations. This electrical signal, which now carries an electrical image of the sound vibrations, is amplified, used to modulate the carrier; the radio wave is radiated, picked up by the receiver, amplified, detected; and finally, as you know, the electrical impulses (similar in shape to those which were originally produced by the microphone) are used to actuate the speaker which in turn recreates the original sound.

Fig. 1-1(B) shows that a system very similar to this is used in television and that in general a very close resemblance exists between sound broadcasting and television. To simplify the explanation at the present time, we assume that only a very small part or element of a scene is being televised. For example, we might allow the light from a neon tube operated from the 60-cycle power line to fall on a small piece of ground glass. The illumination on the ground glass would change from dark through various shades of brightness and back again to dark, and repeat this cycle 120 times per second. (Note that the rate is 120 cycles because both positive and negative cycles cause the neon tube to illuminate the screen.) In the same way that sound broadcasting uses a *microphone* to convert the sound pressure variations into electrical variations, so the heart of any television broadcasting system is the *"television camera"* which converts the time variations in illumination of the scene into corresponding electrical variations. In the simple illustration we have chosen, because only a single small area is being televised, an ordinary photoelectric cell can serve as the television camera.

Once the varying light values have been changed into corresponding electrical values by the television camera, the process of transmitting the information follows exactly the same procedure as in the case of sound broadcasting. Note that the carrier is modulated in the same way, and that it remains stationary in amplitude during the period before the screen is illuminated. Once the neon tube is turned on and illuminates the screen, the amplitude of the carrier varies in proportion to the amount of illumination. Note that the maximum amplitude of the carrier corresponds to a black image, and that the image gets progressively lighter as the amplitude of the carrier is decreased. This is called negative modulation about which we shall have a great deal more to say later on.

For the present, the important thing to note is the similarity between the two systems, the one for transmitting information on light values, and the other for transmitting information on sound values. At the output of the television receiver, we of course have an important change. Whereas the output of the sound receiver is a speaker which converts the electrical impulses into corresponding sound impulses, the output of the television receiver is a *"picture tube"* or other device which converts the electrical impulses into corresponding light values.

We see then, that the a-m sound system and the television system are identical with the exception that the television camera is substituted for the microphone and the picture tube for the loudspeaker. We might also mention here that in the RCA television system, the trade-mark name "Iconoscope" is used for the television camera tube and the trademark name "Kinescope" is used for the picture tube. As will be explained in detail later on, the Iconoscope consists of a very large number of minute photoelectric cells which create an *electrical* picture of the scene being televised, while the Kinescope consists of a cathode-ray tube, on the screen of which is built up a visible image.

Scanning

No doubt you have noticed by this time that in comparing television with a-m sound broadcasting we limited ourselves to televising the simplest type of object, one which was uniformly illuminated over its entire area. We then showed that the two systems are identical provided that the television camera replaces the microphone and the picture tube replaces the loudspeaker. Unfortunately, however, television is not as simple as this or television would have "arrived" many years ago. In television, we are confronted with the problem of conveying information on the light value not at one point but at every point over the complete area of the scene being televised. Thus, the scene must be broken up into a great many elements or elemental areas and information on the light values over each one of these elements must be conveyed to the receiver and finally to the picture tube. And not only must this information be conveyed, but it must be reassembled in the correct order at the receiver and the corresponding light value reproduced for every one of the many elements into which the image has been broken down.

As a matter of fact, this process of breaking down a picture into a great number of elements is nothing new but is as old and as fundamental as the process of seeing itself. For, in viewing a scene, the image is carried to the brain by the eye over a huge network of transmission lines which tells the brain the intensity and the color of the light at every point in the field of vision. Because the number of elements into which the retina of the eye breaks down the scene is so great, we are not conscious that the picture is made up in this way but receive the impression that the picture is perfectly blended or continuous.

Some of you will be surprised to know that even photographs are made up of elementary particles even though they too appear to be continuous upon casual inspection. Actually the light and dark parts of a photograph are the result of the presence of black particles of silver which vary in number over the area of the picture. Where the picture is dark, these particles of silver are more numerous than where the picture is light. Because these particles are so small they are not ordinarily visible. We might note in passing that where photographs are to be enlarged appreciably so-called fine-grain film and special developers are used so that the individual grains or particles will not become visible.

Number of Elements Required

It is important for an understanding of television to appreciate how the number of particles or areas into which a picture is broken up affects the type and quality of the reproduction which is obtained. As we should expect from the preceding discussion, the quality of the picture will be improved as the number of elements is increased. In order to compare the effect of breaking up a picture into a varying number of elements, let us consider the two reproductions shown in Fig. 1-2(A) and Fig. 1-2 (B). These are reproductions of the same photograph, the only difference being the number of elements into which the picture is divided. If you examine these figures closely, you will see that they are composed of a large number of black dots of different sizes, and that Fig. 1-2(B) contains a larger number of dots than Fig. 1-2(A). In printing as well as in the processes of seeing and photography, it is necessary for the picture to be broken down into a series of small areas before it can be printed. The engraver in making his halftone, places a fine mesh screen in front of his camera so that the image is broken down into a series of dots, the actual size of any one dot depending upon the amount of light on the area which the dot represents. Where the picture is dark, the size of the corresponding dot is large, and where the image is light the size of the black dot is correspondingly small. The number of dots into which Fig. 1-2(A) is broken down is 50 per inch, while Fig. 1-2(B) is broken down into 100 per inch.

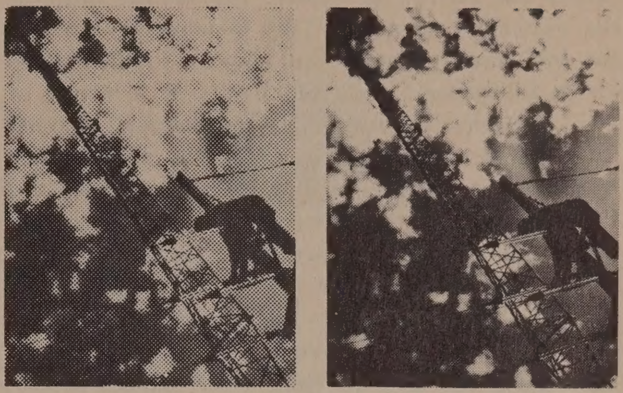


Fig. 1-2.—Halftone reproduction of a photograph using a screen of 50 dots to the inch is shown in (A), at left. The greater detail in (B), at right, was obtained by the use of a screen of 100 dots to the inch.

Courtesy Westinghouse

The great improvement effected by breaking down the picture into a larger number of dots or elements is clearly apparent in the superiority of Fig. 1-2(B) over Fig. 1-2(A). Because of the larger number of elements, the former presents more detail, appears finer, better blended and more continuous than the picture with the fewer number of elements.

The actual number of elements into which a picture must be divided depends upon several factors: the fineness of detail which it is desired to reproduce, the distance from which the picture is viewed, and the size of the picture. Fineness of detail, as we have seen, requires a large number of elements for a given portion of a scene. In addition, the more closely the picture is viewed, the smaller must be the individual elements. This is necessary so that the individual elements will appear to the eye to merge into smooth lines and shades. If a picture of a given scene is enlarged, either the total number of elements must be increased, or the picture must be viewed from a greater distance. Figs. 1-2(A) and 1-2(B) illustrate the first two factors, fineness of detail and viewing distance. Fig. 1-3 which shows Fig. 1-2(B) *enlarged* to twice its original size, illustrates the third factor. Fig. 1-2(B) contains 100 dots per inch; Fig. 1-3 contains only 50 dots per inch. Each contains the same *total* number of elements, therefore the *fineness of detail* is the same in each. Because Fig. 1-3 has *larger* dots, it must be viewed from a greater distance. This consideration is important in television, since the total number of lines into which the scene is divided is the same at all times. Therefore, large pictures should be viewed from a greater distance than small ones to get the same effect.

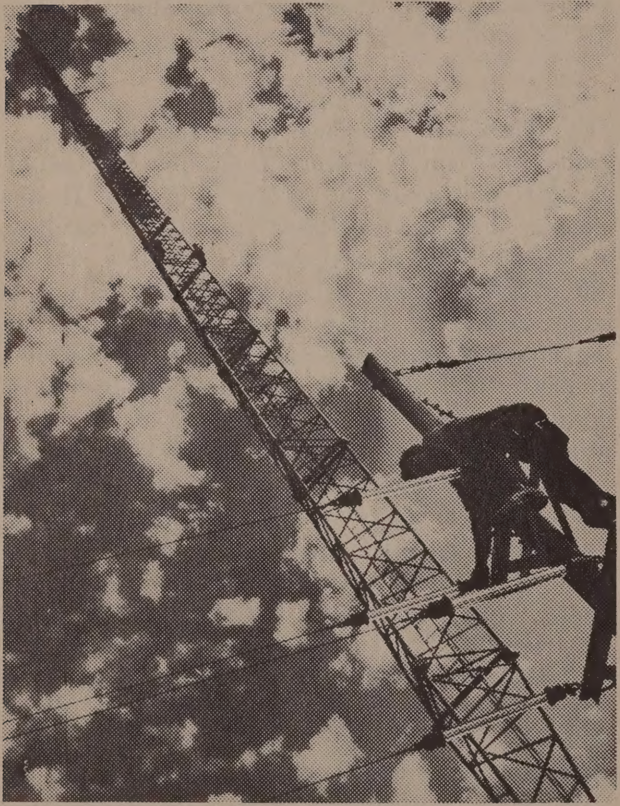


Fig. 1-3.—The same detail as in Fig. 1-2(B) can be obtained by viewing this enlargement from twice the distance. *Courtesy Westinghouse*

An idea of the number of elements needed to reproduce fine details can be obtained from Fig. 1-2 (B). This has 100 dots or "elements" per inch in an area 1.5 by 2 inches, giving 150 × 200 = 30,000 total elements, or 10,000 elements per square inch. Television pictures may be considered to contain about 224,000 elements, regardless of picture size. Although this does not work out to be so great a number per square inch on large picture-tube screens, the fact that the scene is usually in motion compensates for some loss of detail. A point to remember about television is that increasing the number of elements increases the frequency bandwidth which must be transmitted. Thus there are technical and economic limitations to the degree of detail that may be provided.

We can now begin to appreciate the complexity of the problem with which television is confronted. For not only must information on the light value of each one of many thousands of elements be transmitted to the receiver, but also information as to the order or sequence in which these light values must be assembled to form the picture. To make the problem even more complex, all this information must be transmitted in approximately 1/30th of a second in order to prevent blurring due to movement in the scene and in order to make way for the next picture impression or "frame." In this respect the problem of television is more difficult than that of facismile, since in the latter a still picture is transmitted and the time consumed may be ten minutes or more instead of 1/30th second.

Need for Scanning

By this time we have seen that to make television possible, the picture must be broken down into a large number of elements and information transmitted on the light value at each one of the elements. At the receiver this information is reassembled in the proper space relationship to form the original picture. How to transmit this information is the next problem.

Previously we saw that using a conventional a-m system of radio we could transmit information on the light value at any *one* element of a scene by using a photoelectric cell pickup to convert the light value to an electrical value, and that this process was essentially the same as that of sound broadcasting and involved essentially the same transmitting and receiving equipment. This was shown and explained in Fig. 1-1. The first thought that arises is this:

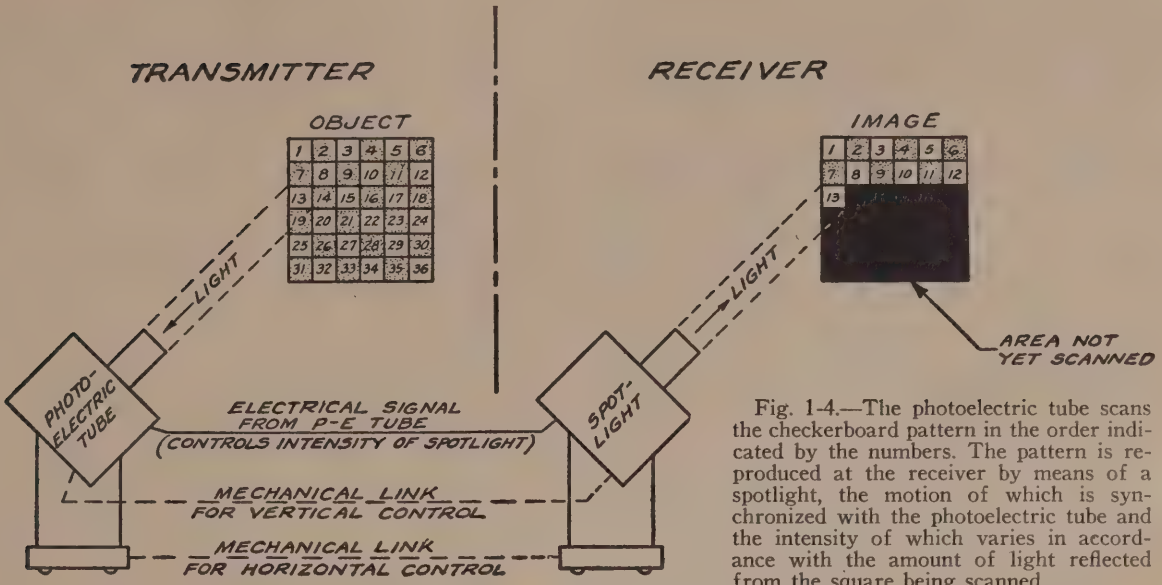


Fig. 1-4.—The photoelectric tube scans the checkerboard pattern in the order indicated by the numbers. The pattern is reproduced at the receiver by means of a spotlight, the motion of which is synchronized with the photoelectric tube and the intensity of which varies in accordance with the amount of light reflected from the square being scanned.

Why not use individual channels of the type shown in Fig. 1-1(B) to transmit information on the light values at each one of the points into which the picture is divided? But then, this would require some 100,000 individual pickups and transmission systems each of which would be similar to the system of Fig. 1-1(B) Obviously such a system would be far too complex and expensive to be practicable, even if other problems of great difficulty at the receiving end did not exist.

A more promising solution, which is used in all television systems today, works on the basic idea of transmitting information on the light value of one element at a time. In this way the picture is covered or "scanned" in a systematic way until finally the image is said to be completely scanned when information on the light values at every point in the picture has been obtained. The need for more than one channel is thus avoided. However, the information must still be reassembled in the proper order at the receiving end before the picture can be obtained.

Scanning a scene is sometimes compared with the manner in which we read a printed page. For instance as you read this page your eyes start at the upper left-hand corner of the page and successively sweep to the right-hand side while examining every letter on the line; when the first line has been completed the eye snaps over rapidly to the left-hand side, jumps down one line, and in the same way, examines every letter on the second line as it progresses at a uniform rate toward the right. This procedure continues until finally the eye reaches the last letter on the lower right-hand side of the page, at which time we can consider that the whole page has been scanned.

This same procedure is used when a television camera scans a scene which is to be televised. The only difference is that the television camera breaks down the scene into finer elements than the letters of a page and that the camera produces electrical impulses which vary in proportion to the amount of light on each element of the scene being scanned.

The following simple example will help to clarify the fundamental principles and requirements involved in scanning. Suppose we wish to reproduce by television the pattern shown at the left of Fig. 1-4. Let us assume in this example that a television camera (consisting essentially of a photoelectric cell with the proper lens equipment) is available which can be focused so as to pick off the light values on any one of the squares into which the picture has been divided. Let us also assume that we have a mechanical arrangement for moving the camera both horizontally and vertically so that it will scan the object. That is, it is possible to start with the camera focused on element 1 and to move it at a uniform rate across the screen until it reaches element 6. At this point the camera snaps back rapidly to element 7 at the beginning of the second line and moves at a uniform rate along the second line until it reaches element 12. The camera then returns very rapidly to the left and starting at element 13 on line 3 it scans the third line. In this manner the procedure continues until the entire pattern is scanned. Note that at any one instant the camera receives light only from the particular element at which it is aimed and focused and produces an electrical impulse which is proportional to the amount of light reflected by this element.

So much for the scanning at the transmitting end where the picture is being televised. At the receiving end let us assume that we have a projector which projects a narrow pencil of light on the screen equal in area to one of the square elements. This projector like the camera can be moved horizontally and vertically so that the light can be focused on any part of the screen. Suppose further that the electrical impulses from the television camera are fed to the projector and arranged to control the intensity of the light emitted by the projector in accordance with the amount of light registered by the camera at any particular instant.

Under these conditions before a picture can be obtained at the receiver, the motion of the camera at the scene being televised and the motion of the projector at the receiver must be properly coordinated or "synchronized." This means that the television camera and the projector must go through the same movements together, that the projector must at all times be focused on exactly the same element in the picture as that on which the camera is focused. In the figure we have assumed a sort of mechanical linkage between the camera and the projector to accomplish this; actually no such mechanical linkage is possible in television and we shall see later that electrical *synchronizing pulses* are used to control the camera at the transmitting end and the projector (or picture tube) at the receiver, so that both the scene being televised and the image which is being reproduced at the receiver are scanned in unison—so that the scanning is *synchronized.*

In the picture shown, the image has been scanned only as far as element 13; element 14 is about to be scanned. As a result the image at the receiver is totally dark beyond this point since the lower elements have not yet been scanned and hence have not yet been illuminated. We shall explain later on that the observer sees the complete image at one time even though only one element of it is receiving light at any particular instant. This is because the entire scanning process is repeated some thirty or more times a second, and the eye tends to see the image after it is no longer illuminated.

We can now summarize the requirements which must be met before a scene can be transmitted by television:

1. The scene must be systematically scanned by the television camera which interprets the light values at every element of the scene in terms of corresponding electrical values.

2. The image must be scanned at the receiving end according to the same systematic plan used by the television camera and the intensity of the light emitted by the light source in tracing the image must vary at every instant in accordance with the amount of light which the camera is receiving at that instant.

3. At every instant the camera and the light tracing the image must be synchronized so that the identical portion of the image is being traced out which corresponds to the element of area being scanned by the camera.

4. This scanning procedure or process must be completed over and over again at a rate of at least 30 times per second so that as far as can be determined by the eye a continuous image of the scene is formed.

THE CAMERA AND PICTURE TUBES

We have already seen that a complete television system, like a complete broadcasting system, requires a pickup at the transmitter and a reproducer at the receiver, and that the pickup is a photoelectric tube and the reproducer is a cathode-ray tube similar to those used in oscilloscopes. To avoid confusion of trade names let us call the pickup a "camera tube" (it "takes" the television picture) and the reproducer a "picture tube" (it reproduces the picture). Since we want to get a good general idea of how a television system works, we shall consider both these tubes here, although the serviceman in the field will naturally come in contact with only the picture tube.

The Picture Tube

Essentially the television picture tube is similar to the familiar cathode-ray oscilloscope tube, so that those who have read Rider's "The Cathode-Ray Tube at Work" will have a good basis for under-

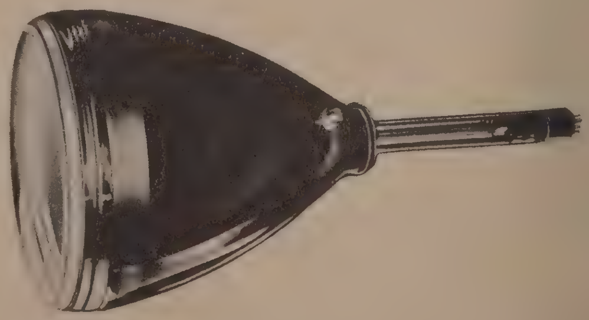


Fig. 1-5.—A Kinescope (picture tube) with a 12-inch screen.

Courtesy RCA

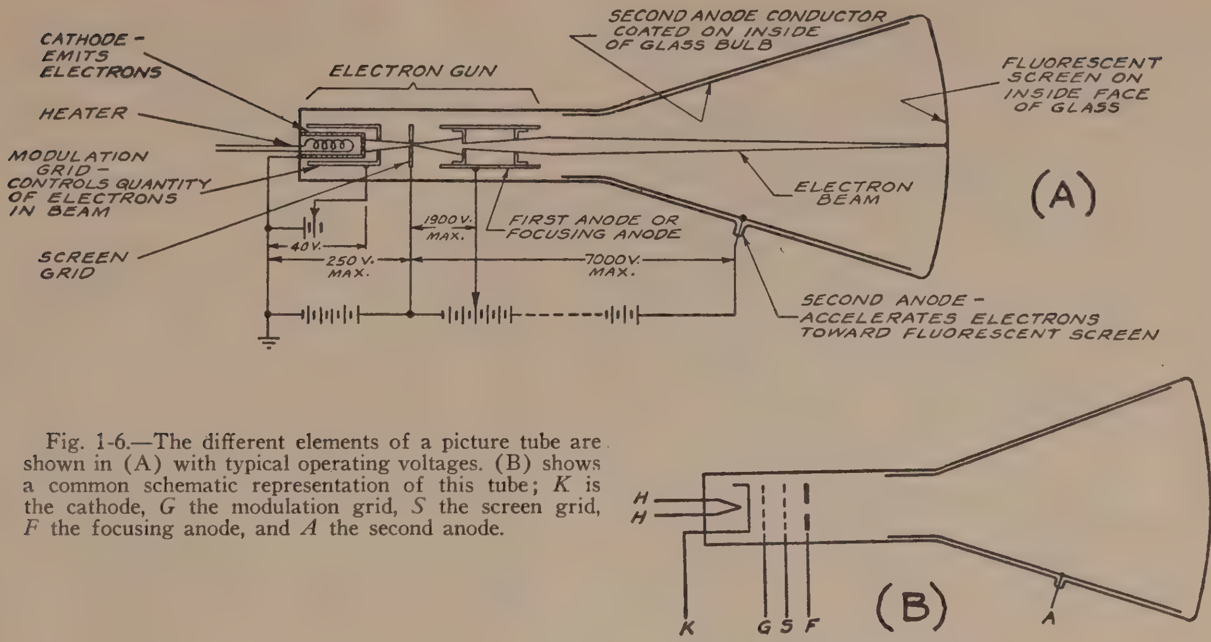

CATHODE - EMITS ELECTRONS
HEATER
MODULATION GRID - CONTROLS QUANTITY OF ELECTRONS IN BEAM
SCREEN GRID
ELECTRON GUN
SECOND ANODE CONDUCTOR COATED ON INSIDE OF GLASS BULB
FLUORESCENT SCREEN ON INSIDE FACE OF GLASS
ELECTRON BEAM
(A)
1900 V. MAX.
FIRST ANODE OR FOCUSING ANODE
40 V.
250 V. MAX.
7000 V. MAX.
SECOND ANODE - ACCELERATES ELECTRONS TOWARD FLUORESCENT SCREEN
H
H
K
G
S
F
(B)
A

Fig. 1-6.—The different elements of a picture tube are shown in (A) with typical operating voltages. (B) shows a common schematic representation of this tube; *K* is the cathode, *G* the modulation grid, *S* the screen grid, *F* the focusing anode, and *A* the second anode.

standing television picture tubes. For those who have not we will review the subject here.

Let us assume that television (video) signals are coming into a receiver; as we have said before, the amplitude of these signals is proportional to the light reflected by the object being televised. We want to use these signals to produce a picture. In sound work we know that the signals can be made to move the diaphragm of a loudspeaker, thus producing sound waves similar to the original. The picture tube, then, must be capable of converting the electrical video signals into *light* to produce a picture. It is a property of certain substances called fluorescent materials that they will glow when they are struck by a beam of electrons, and the more electrons striking such a substance at a given instant, the brighter will be the glow. A picture tube, then, can be made if we have a source of electrons, means for controlling their motion and their quantity, and a fluorescent screen.

The external appearance of a typical picture tube is shown by the photograph, Fig. 1-5. This is a glass vacuum tube specially shaped to withstand the high pressure exerted by the surrounding air, due to the high vacuum within the glass envelope. Servicemen should remember this, and handle picture tubes with great care. Even scratching the glass or careless handling may cause them to *collapse* as violently as if they exploded. The white appearance of the large end of these tubes is caused by the film of fluorescent material deposited on the inside surface; this, of course, is where the television pictures are formed.

A cross-sectional view of a picture tube is shown in Fig. 1-6. As in the usual radio tube, the heater causes the cathode to emit electrons and the second anode (like the plate of the ordinary tube) strongly attracts them, giving them a high velocity. The modulation or control grid regulates the number of electrons which pass through it in a given time. In the picture tube additional elements such as a focusing anode form the electrons into a narrow beam so that they will strike the fluorescent screen in a small round spot; in some tubes a screen grid is inserted between the modulation grid and the focusing anode to prevent the focusing action from affecting the modulating action. Fig. 1-6(A) shows the general arrangement of parts inside the tube, and Fig. 1-6(B) shows a common way of representing these in schematic form.

With the parts so far mentioned, the tube can produce a beam of electrons which will hit the center of the fluorescent coating on the inside of the picture tube and produce a small spot of light which can be seen through the glass end. We can focus this spot by varying the potential on the first (focusing) anode, and we can vary its brightness by applying a suitable potential to the modulation grid. The more negative the modulation grid becomes relative to the cathode, the dimmer the spot becomes; the less negative the grid, the brighter the spot. It now remains to provide some means for moving this spot rapidly enough over the fluorescent screen to give us a complete picture . . . in other words to provide scanning.

Two methods of deflecting the electron beam are now commonly used: *electrostatic* deflection and *electromagnetic* deflection. The first of these methods, which is probably the simpler to understand, takes advantage of the familiar fact that particles of matter having like charges of electricity repel each other, while particles having unlike charges attract each other. Since the electron beam consists of negative charges, we see immediately that the beam can be deflected by means of suitably shaped electrodes which are charged either positively or negatively as required. In picture tubes, as in oscilloscope tubes, this is done by building into the tube two pairs of metallic plates, arranged approximately as shown in Fig. 1-7.

This figure shows that plates H_1 and H_2 are parallel to each other but in a plane at right angles to plates V_1 and V_2. If no potential is applied to any of these plates, the electron beam will pass straight along the axis of the tube and cause a spot to appear on the screen at A. Now if we leave plates H_1 and H_2 alone, but make plate V_1 positive with respect to plate V_2, plate V_1 will tend to attract the negative electrons which make up the beam, thus causing the beam to bend, so that it strikes the screen at a new point, say at point B. We have thus deflected the beam upward in a vertical direction a distance AB. Similarly, if we leave V_1 and V_2 alone, but make H_1 positive with respect to H_2, H_1 will deflect the beam horizontally to the right until we get a spot at a point such as C.

If we combine these effects, by making both V_1 and H_1 positive at the same time, the beam will bend sidewise and upwards, causing the spot to appear at D. If the potentials of V_1 and H_1 have been the same as used in the first two tests, D will be located so that the distance AD is the hypotenuse of a right triangle whose sides are equal to DC and AC. Naturally, if V_2 is made positive with respect to V_1, the beam will be deflected vertically downward, and if H_2 is made positive with respect to H_1, the beam will be deflected horizontally to the left. The whole process of electrostatic scanning is simply a matter of varying the potentials on the deflecting plates in the picture tube so that the spot on the screen traces out the desired picture according to a regular plan. In a later section, we shall study the details of the plan actually used.

The second method of deflecting the electron beam is called *electromagnetic deflection,* because coils of wire carrying current act on the beam the way a magnet would. See Fig. 1-8(A): The magnetic lines of flux from a permanent magnet pass from the north pole (N) to the south pole (S); the electron beam passes between the poles. If the magnet were not present, the beam would produce a spot at point A on the screen. When the magnetic field acts on the beam as in Fig. 1-8(A), however, the beam is deflected vertically upward to produce a spot at point B. Compare this effect with that produced by the electrostatic field in Fig. 1-7, and you will notice that in Fig. 1-7 the electron beam was deflected *in the direction of the lines of electrostatic flux* existing between plates V_1 and V_2, whereas in the case of magnetic deflection, the beam is deflected *at right angles to the lines of magnetic flux.* Thus in Fig. 1-8(B) the beam is deflected horizontally to the right, when another magnet is introduced whose influence is at right angles to the one shown in Fig. 1-8(A).

Of course, a permanent magnet gives a steady

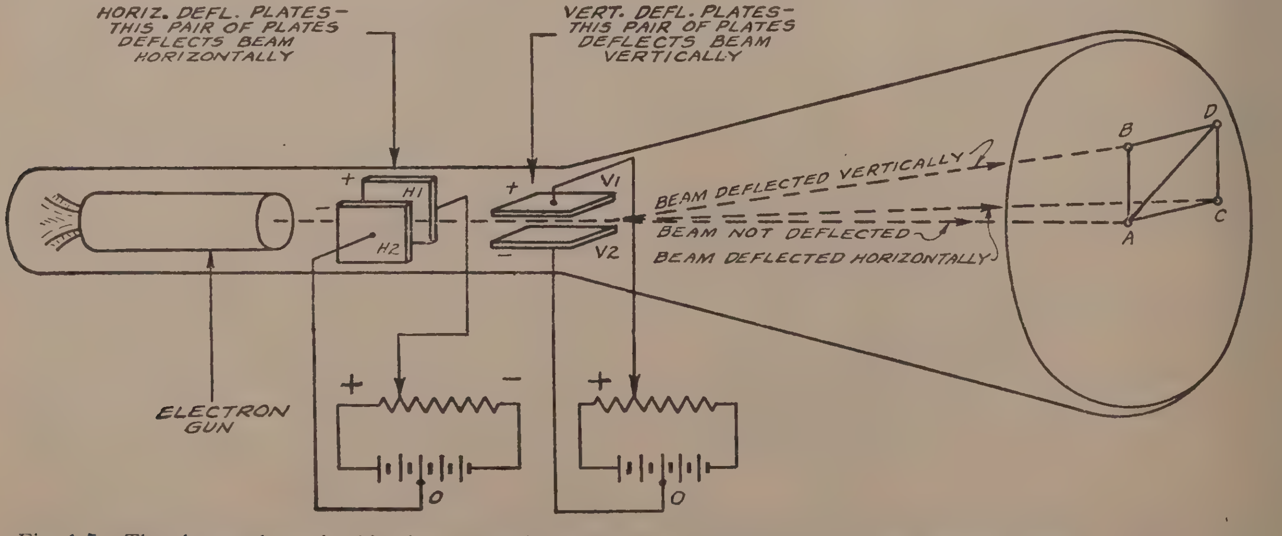

HORIZ. DEFL. PLATES– THIS PAIR OF PLATES DEFLECTS BEAM HORIZONTALLY
VERT. DEFL. PLATES– THIS PAIR OF PLATES DEFLECTS BEAM VERTICALLY
+
H1
H2
+
V1
V2
–
BEAM DEFLECTED VERTICALLY
BEAM NOT DEFLECTED
BEAM DEFLECTED HORIZONTALLY
B
D
A
C
ELECTRON GUN
+
–
+
O
O

Fig. 1-7.—The electron beam in this picture tube is electrostatically deflected by the two pairs of plates. H_1 and H_2 deflect the beam horizontally, and V_1 and V_2 deflect it vertically.

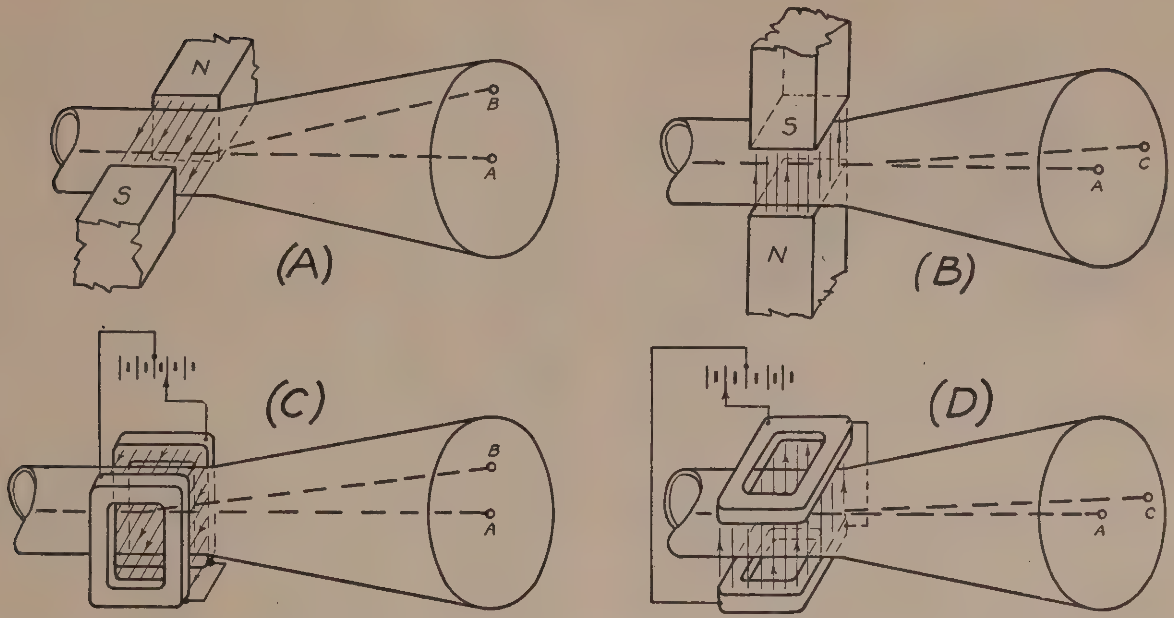


Fig. 1-8.—The magnetic deflection coils in (C) and (D) provide a magnetic field similar to that of the permanent magnets in (A) and (B). Note that the beam is deflected at right angles to the lines of force.

field and would therefore produce a constant deflection, whereas in television we must be able to vary the extent of the deflection from zero to maximum in various directions. Since we can also produce a magnetic field by passing a current through a coil, and in addition, can vary the strength of the field by varying the current, we use coils for electromagnetic deflection, as shown in Figs. 1-8(C) and 1-8(D). Two pairs of coils are used, and these are often combined in a single compact cylindrical unit called a "deflecting yoke." Fig. 1-9 shows an RCA deflecting yoke.

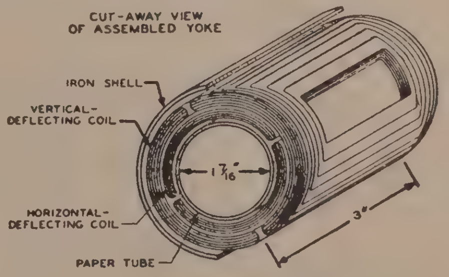
A point to remember about deflection systems is that when one tube has *electrostactic* deflection, a change of voltage on the deflecting plates is required to move the beam, whereas in *electromagnetic* deflection, a change of current through the deflecting coils is required.

Recently it was discovered that a metallic film placed over the luminescent material on the inner surface of the Kinescope tube improved the distribution of light over the screen. This metallic film is in the form of an aluminum backing which acts as a reflector. In operation, the electrons pass through the aluminum and strike the luminescent material causing it to fluoresce in the usual manner. However, very little light is lost within the tube because of the reflecting properties of the aluminum surface. The result is a gain in the brightness of the received image. It is probable that this new type of Kinescope may be used quite extensively in new models of television receivers.

The Camera Tube

Whereas the picture tube contains a fluorescent screen to convert an electric current into light, the camera tube contains a photosensitive screen to convert light into an electric current. Large and small RCA camera tubes, called Iconoscopes, the type in general use in this country at the present time, are illustrated in Fig. 1-10. The essential parts of the large tube are shown in the cross-sectional view in Fig. 1-11: these are the mosaic, the signal plate, the collector, and the electron gun. The important accessories external to the tube are the lens, the deflecting yoke, and the load resistor.

The mosaic consists of millions of individual



Courtesy RCA

Fig. 1-9.—The magnetic deflecting yoke contains the horizontal and vertical deflection coils which are designated here.

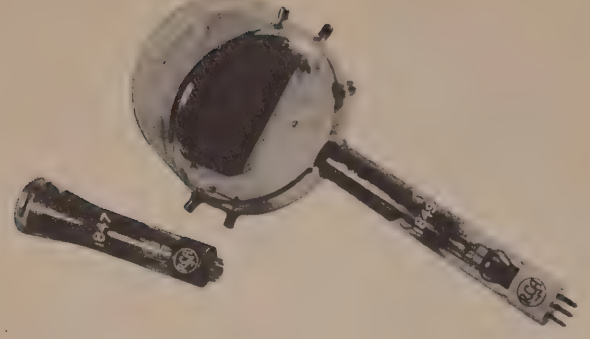


Fig. 1-10.—Both a large and a small Iconoscope are shown.
Courtesy RCA

photosensitive globules like metallic droplets deposited on one side of a thin sheet of mica. Each globule is like a minute island on the mica, so that the globules are insulated from each other. On the other side of the mica there is a layer of conducting material . . . the signal plate. Because the globules are separated from the signal plate by the mica, the mosaic consists of a myriad of mica-dielectric capacitors, all having one plate in common. Fig. 1-12 shows in a general way what the mosaic looks like when viewed from the edge and enormously magnified. The collector ring is a metallic coating applied in the form of a ring around the inside of the tube and also extended down the neck of the tube near the electron gun.

Briefly the action of the Iconoscope is as follows: An optical image of the scene to be televised is focused on the mosaic by the lens. The electron gun projects a stream of electrons over the mosaic, scanning it under control of the deflecting yoke, as already described in connection with the picture tube. The electron beam in traversing the mosaic causes a succession of voltages to appear on the signal plate which are proportional to the distribution of light in the image of the scene. The resultant current which flows in the load resistor causes a voltage drop. This constitutes the *video signal.* This signal is then amplified and used to modulate the television transmitter.

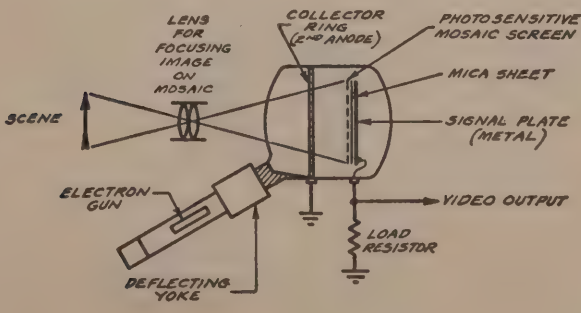
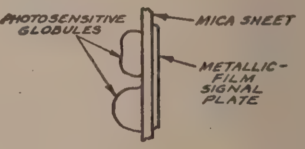

LENS FOR FOCUSING IMAGE ON MOSAIC
COLLECTOR RING (2ND ANODE)
PHOTOSENSITIVE MOSAIC SCREEN
MICA SHEET
SCENE
SIGNAL PLATE (METAL)
ELECTRON GUN
VIDEO OUTPUT
LOAD RESISTOR
DEFLECTING YOKE

Fig. 1-11.—The elements of the large tube shown in Fig. 1-10 are easily identified. Light images formed on the signal plate are transformed into electrical energy by the scanning beam.

The following more detailed description of the action of the Iconoscope is presented for those interested. When a scene is focused on the mosaic, each photosensitive globule emits electrons in proportion to the amount of light falling upon it; the more light that falls on a given area the more electrons are emitted from that area. This process is called "photoemission." Since a loss of electrons means a more positive charge, photoemission sets up a variety of different charges over the surface of the mosaic in accordance with the distribution of the light it is receiving. So far no video signal has been produced because the globules are insulated from the signal plate.

Since the path from the mosaic to the signal plate is through a capacitor, the only way to get a signal across is to produce a sudden *change* of potential. In the Iconoscope type of camera tube, this change of potential is produced by the action of the electron beam from the gun. (The necessary orderly scanning action from left to right and top to bottom is provided by suitable currents through the deflecting yoke, as in the case of a magnetically deflected picture tube.) In order to understand the effects produced by this scanning beam, let us first consider three possible conditions on the surface of the mosaic *before* scanning. It has been found that the ordinary Iconoscope mosaic in *total darkness* (black scene) asumes a potential of —1.5 volts relative to the collector ring. When strong light (white scene) falls on a portion of the mosaic, a considerable number of electrons are lost by photoemission, and the potential is changed from, say, —1.5 volts to 0 volt. Illumination of medium intensity (gray scene) may cause the potential to change from —1.5 volts to —0.8 volt. These potentials of the mosaic, then, may be approximately as follows for these portions of the mosaic: black area, —1.5 volts; gray area, —0.8 volt; white area, 0 volt.

Fig. 1-12.—An enlarged view of the way the photosensitive mosaic is arranged with respect to the signal plate of a camera tube.



The electrons in the scanning beam travel so rapidly that when they strike a surface such as the Iconoscope mosaic, they knock off a great many electrons, more, in fact, than even bright light does. This kind of action is called "secondary emission." So many electrons are thus lost by secondary emission that the portion of the mosaic directly under the action of the scanning beam is driven to a *positive potential* of +3 volts. Each portion of the mosaic is driven to +3 volts *during the instant the scanning beam acts upon it* and this +3-volt potential is reached *regardless of the light conditions prevailing* . . . a black area goes to +3 volts as well as a white or gray area. Although the *peak* potential reached by every area as the scanning beam passes over it is the same, the *change in potential* which this area undergoes at this time *will depend on the illumination* it has received. Thus the black area will change from —1.5 to +3 volts, a change of 4.5 volts; the gray area will change from —0.8 to +3 volts, a change of 3.8 volts; the white area will change from 0 to 3 volts, a change of 3 volts. It is this sudden *change of potential* which is induced on the signal plate that causes the video signal current to flow through the load resistor. The *difference between the changes* for black, gray, and white areas is what indicates the difference in the illumination over the mosaic surface as it scanned, and this difference is proportional to the distribution of light over the mosaic.

The important things to notice here are: (1) that we get a video signal across the load resistor (Fig. 1-11) only when a *change of potential* is produced on the mosaic, and that this is accomplished by the scanning beam; and (2) that the resultant *video signal is proportional to the light coming from the scene being televised,* the signal being of *maximum* amplitude for *black* portions of the scene, *minimum* amplitude for *white* portions.

The collector ring (2nd anode) serves to accelerate the electrons in the scanning beam, and also to collect some of the electrons emitted from the mosaic. In Fig. 1-11 you will notice that this element is grounded; it is, however, 500 to 1000 volts *positive* with respect to the cathode of the electron gun, because the gun is at a corresponding potential *negative* with respect to ground. This will not be the last time that you will find television tubes with elements at very high negative potentials, and the possibility of such circuits existing must always be borne in mind as a safety measure.

Another type of RCA camera tube more recently developed is the Image Orthicon. This tube retains the essential features of the Iconoscope but it is much more sensitive and efficient. Its high degree of sensitivity permits the pickup of scenes in candlelight and varying degrees of darkness, and makes possible "round-the-clock" television coverage of news and special events. This tube, shown in Fig. 1-13(A), is incorporated in the portable super-sensitive television camera manufactured by RCA Victor Division.

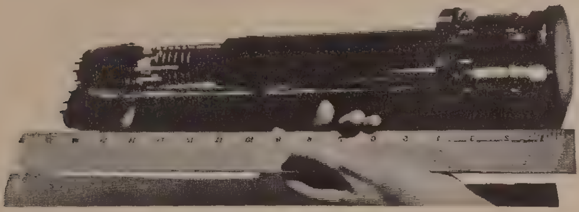
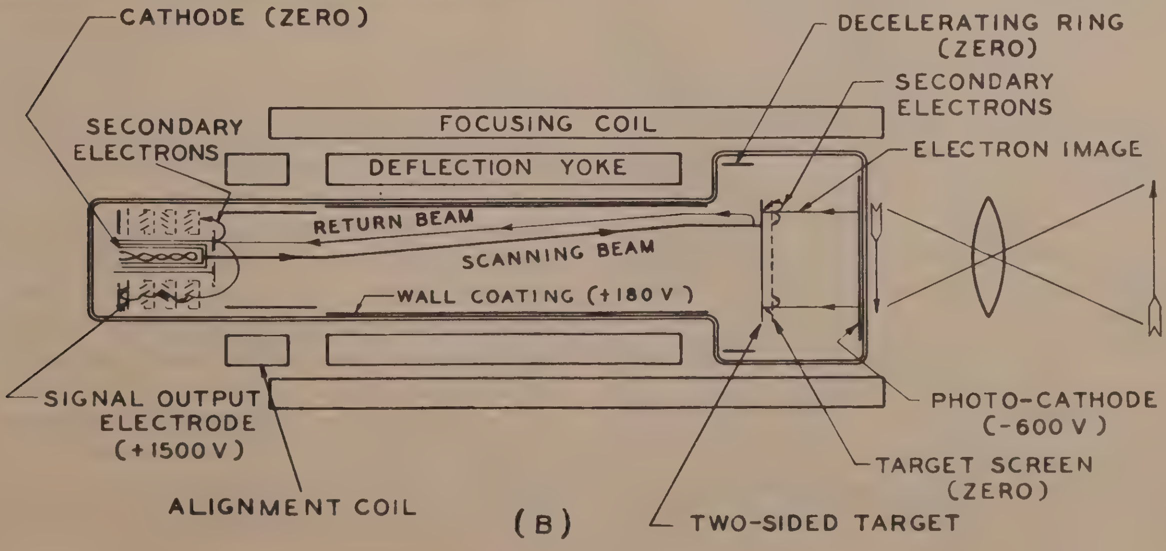


Fig. 1-13.—An Image Orthicon is shown in Fig. 1-13(A). Note the size of the tube. The elements of the tube (including the deflection yoke and focusing coil) are shown in Fig. 1-13(B).

Courtesy RCA



A simplified cross-section of the Image Orthicon is shown in Fig. 1-13(B). In operation, the scanning beam is projected from the cathode (extreme left) toward the target at the right on which electrical charges have been built up for focusing an image of the scene on the photo-cathode. The scanning beam is given just enough velocity to reach the target. If the electrons in the beam strike a positive charge they are absorbed. If the target charge is negative, electrons in the beam are returned to the area near the cathode. Incorporated within the tube near the cathode are reflecting electrodes known as dynodes. These dynodes are coated with a material capable of high secondary emision. Referring to Fig. 1-13(B), when the electrons in the return beam strike the first of the dynodes, secondary electrons are emitted. These secondary electrons in turn strike other dynodes, greatly increasing the total number of electrons and hence the current carrying the picture information. This procedure is repeated for each element in the televised scene.

SCANNING AND SYNCHRONIZATION

In previous sections we discussed (1) the necessity for scanning both the camera tube and the picture tube, (2) the essential requirement that these operations must be synchronized, and (3) the devices which make it possible to control the motion of the electron beam in these tubes. In this section we shall describe the details of the path of the scanning beam on the tube screen, the waveshapes necessary to produce this path, and some of the problems which arise in connection with scanning.

The Scanning Pattern

As we have already mentioned, the usual scan is from left to right and from top to bottom. In certain cases this order must be reversed, but the principle is the same. In Fig. 1-14 is shown a simplified diagram of a scanning pattern; this is often called a "raster." The rectangular diagram formed by the light lines connecting points *A, B, C,* and *D* represents what we may call the "picture space." The width of this rectangle (*AB*) is drawn so as to be 4/3 times the height (*AC*); this is the proportion used in motion pictures. The ratio of width to height is called the "picture aspect ratio" and its value of 4/3 is one of the FCC television standards. On a picture tube, the heavy lines *ab, cd,* etc., are caused by the fluorescent glow which occurs as the electron beam moves from point *a* to point *b* (in the direction of the arrow on line *ab*). You will notice that points *a* and *A* coincide, while point *b* is slightly *below* point *B*. In fact, *all* the lines in Fig. 1-14 slant downward. This slant is necessary to provide the *vertical* (top-to-bottom) part of the scanning process. In fact, the left-to-right component of scanning is called "horizontal scanning" or "line scanning"; the top-to-bottom component is called "vertical scanning" (also called "frame scanning" or "field scanning," for reasons which will be clear later on). This slant-line system of scanning is used because the characteristics of the electrical circuits which provide scanning make it impossible to scan in a series of truly horizontal lines, with an abrupt vertical drop at the end of each line.

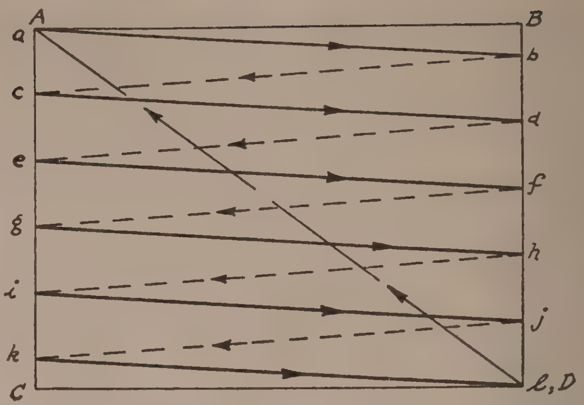


Fig. 1-14.—A simple progressive scanning pattern in which the picture is scanned in a series of lines which start at the upper left and slope down to the right.

After the beam reaches point *b* it is deflected back along the dotted line *bc* to start a new line, *cd*. On the picture tube the line *bc* would be much fainter than line *ab* because the *time* allowed for the beam to move from *b* to *c* is only about one-tenth the time allowed for it to travel from *a* to *b*. Lines like *ab, cd,* etc., are called "line traces"; lines like *bc, de,* etc. are called "line retraces" or "line flybacks."

Here some questions may arise: Why does the motion of the spot across the screen appear as a line? Why is the flyback dimmer than the trace? The eye is responsible for these effects. Any one who has ever swung a flashlight or a "sparkler" in a circle must have noticed that as the speed of the swinging is increased, the individual spot of light merges into an apparently continuous circle of light. The velocity of the scanning spot across the picture-tube screen is so rapid (several thousand miles per hour) that the eye cannot distinguish the spot from a continuous streak. This property of the eye is called "persistence of vision."

The difference in brilliance between the trace and the flyback is due to the fact that the brilliance of the fluorescence produced on the picture-tube screen depends on the *time* during which the electron beam

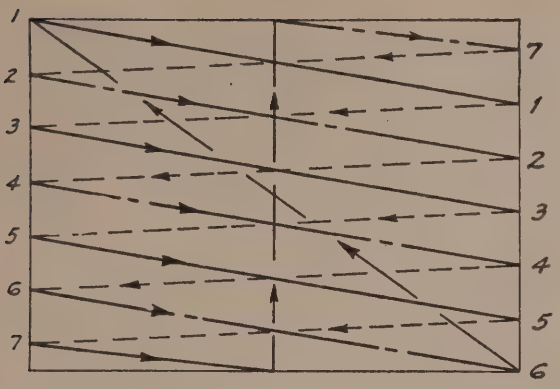


Fig. 1-15.—An interlaced scanning pattern in which the odd and even lines are scanned on successive fields.

acts upon it. Although in both cases the actual time is very short, the *difference* in time (about 10 to 1) does result in a very marked difference in brilliance.

A complete set of lines such as shown in Fig. 1-14 is called a "frame." This particular frame is completed when the spot reaches point 1; it then returns to point *a* to start a new frame. A frame like that shown in Fig. 1-14 is said to be built up by "progressive scanning" (the lines follow each other in a continuous chain). Six complete lines (retraces are not counted) are shown in this figure; a complete technical description of this figure would be: "A six-line frame (or raster) produced by progressive linear scanning." Of course, so few lines as we have shown would be insufficient to give a good picture, and a very great many more are actually used. In order to give the eye the illusion of motion, many complete frames must be traced out every second. The standard American procedure is to trace 30 complete frames every second; each frame consists of 525 lines.

Flicker and Hum on the Raster

It has been found that if the raster is bright a *flicker* will be observed when the number of frames per second (frame frequency) is less than a certain critical value. Although a satisfactory illusion of motion might be produced by about 12 frames per second, as many as 48 frames per second may be required to remove objectionable flicker. The more frames per second a television transmitter sends out, each having a large number of lines, the higher the modulation bandwidth required. One way to keep the bandwidth down is to send twice as many frames per second, each frame having half the number of lines. But a large number of lines is necessary for detail in the picture. It was therefore decided to divide the total number of lines in each frame between two rasters (called "fields"), each containing half the total number of lines required to make up a frame. These fields, transmitted at the rate of 60 per second, get rid of the flicker effect. The required number of lines per frame is supplied by sending the odd-numbered lines along with one field and the even-numbered lines along with the following field. The fields are transmitted so rapidly that as far as the eye can see, all the necessary lines appear in their proper positions on the raster simultaneously.

The method of scanning just described is called "interlaced scanning," and is the system now in use. A simplified pattern of a frame produced by interlaced scanning is shown in Fig. 1-15. The general principles are the same as for progressive scanning; in fact, each *field* is itself progressively scanned. The lines transmitted during the first field scan, referring to Fig. 1-15, are lines 1, 3, 5, and the first half of line 7. The beam is then deflected to the top of the picture space to begin the second field. During the second field, line 7 is completed and lines 2, 4, and 6 are traced, thus completing the entire seven-line frame in two steps. (As in Fig. 1-14, the arrows show the direction of travel of the beam spot.) The use of an odd number of lines has been found advantageous in interlaced scanning; this method is called "odd line interlacing." American practice calls for 262.5 lines in each field, making 525 total lines per frame. Sixty fields are transmitted each second, giving a field frequency of 60 per second.

The frame frequency chosen depends upon the a-c power-line frequency; it must be an even multiple or submultiple thereof. The reason for this is that if there is some hum present in the deflecting circuits, a certain amount of distortion will appear in the picture. If this distortion appears to be stationary, and is not excessive, it can probably be tolerated. On the other hand if the distortion moves so that the picture appears to have moving ripples in it, even slight distortion is most objectionable. By making the frame frequency an even submultiple of the power-line frequency, this type of hum pattern can be rendered stationary. Since most American power lines use 60-cycle a.c., the frame frequency chosen was 30 per second. (In England, with 50-cycle a.c., the frame frequency is 25.)

Scanning Waveform

In describing the picture tube, we showed that the beam can be deflected by applying suitable potentials to the deflecting plates or suitable currents

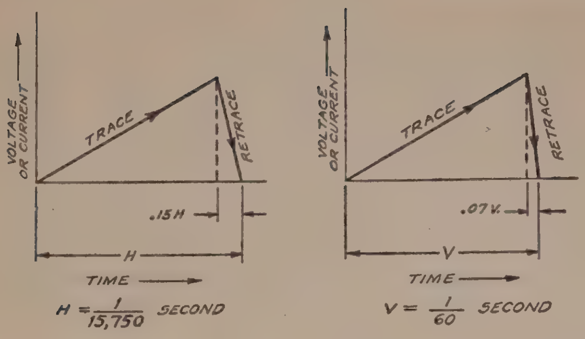


Fig. 1-16.—At the left is shown the saw-tooth waveform required for horizontal scanning. A similar waveform, but of much lower frequency is required for vertical deflection and is shown on the right.

through the deflecting coils. The nature of these voltages and currents depends on the kind of deflection to be produced. In Fig. 1-15 is shown the kind of deflection required in television scanning. The trace deflection required is a steady motion from left to right along a line slanting slightly downward to the right; the retrace travels from right to left along a line slanting downward to the left. We found out that such a motion will occur when both sets of deflecting plates or coils are in operation simultaneously. Thus in scanning, two deflection circuits are acting at once, one moving the beam horizontally (horizontal deflection circuit), the other moving it vertically (vertical deflection circuit).

The shape of the voltage (or current) wave which has to be applied to the deflecting plates (or coils) to produce the desired scan is shown in Fig. 1-16. At the left is shown the waveform for horizontal or line scanning; at the right the waveform for vertical or field scanning. Notice that the time of the line wave is equal to the time of the field wave divided by the number of lines per field (1/60 second divided by 262.5 = 1/15,750 second). Also notice that the line retrace is allowed only about 1/6 the time allowed to the line trace; the field retrace is allowed 1/13 the time of the field trace. The total time intervals allowed for these waves indicate that the line-scan frequency is 15,750 cycles and the field-scan frequency is 60 cycles. These waveforms are called saw-tooth waves; circuits for generating them are described in a later section.

Over-all View of a Television System

We are now in a position to obtain a bird's-eye view of a complete television system. Such a step is advisable at this time because of the complexity of the system and the desirability of not losing sight of the function of the principal parts in the maze of detail associated with all the individual elements of the system.

Fig. 1-17 shows the general role played by each of the major units of a television system in causing the picture of the televised scene to appear on the screen of the picture tube. Referring to this illustration, you will note that the camera tube is focused on the scene to be televised with the result that an image of the scene is formed on the photoelectric mosaic of the camera tube. In this way each point of the mosaic takes on a voltage which is proportional to the light value associated with this point of the image. In order to transmit this information, the electron beam of the camera tube completely scans the image on the photoelectric surface 30 times in each second. As a result of this the video portion of the television signal is produced in which the electrical variations correspond with the light variations on the screen of the camera tube.

In order to fulfill the requirements of synchronization, the need for which has already been explained, a synchronizing signal (abbreviated "sync" signal) is applied to the camera tube deflecting circuits so that the scanning of the electron beam is at all times under the timing control of this sync signal.

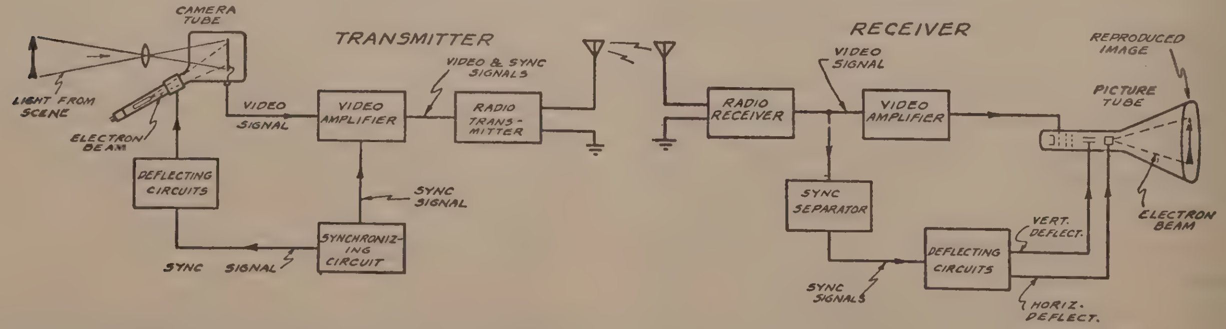


Fig. 1-17.—The principal elements of a complete television system. Note the provision made for synchronizing the scanning at the picture tube with that at the camera tube.

At the same time this sync signal must also form a part of the television signal which is broadcast in order that the scanning at the picture tube in the receiver can be kept in synchronism (in step) with the scanning at the camera tube. For this reason you will note that the sync circuit also feeds the same sync signal to the video amplifier, and as a result the complete signal contains information not only on the light values but also the necessary control signals to synchronize the scanning at the picture tube with that at the camera tube.

The complete television signal is amplified by the video amplifier circuits in the transmitter and fed to the modulating circuits of the transmitter where it modulates the high-frequency radio wave which serves as the carrier. According to the present frequency allocations assigned by the Federal Communications Commission, the carrier frequencies used lie within the range between 44 and 88, and between 174 and 216 megacycles. The reasons for the use of carrier frequencies in the ultrahigh frequency range and the actual make-up of the signal will be discussed later in detail.

The signal which is radiated by the transmitting antenna is picked up by the receiving antenna and fed to the television receiver. The television signal is amplified in this receiver which is almost invariably of the superheterodyne type. After sufficient amplification the signal is demodulated in the second detector of the receiver and the video and sync signals are recovered. The video amplifier following the detector further amplifies the signal which is finally impressed on the control grid of the picture tube. The fluctuations of voltage on the control grid of the picture tube cause the intensity or brightness of the scanning spot to vary in accordance with the amount of light on that element in the scene which at that particular instant is being scanned by the electron beam of the camera tube.

The receiver contains separate circuits for deflecting the beam of electrons horizontally and vertically so as to accomplish the scanning of the image at the picture tube. In general these circuits are similar to those used to deflect the electron beam at the camera tube. To insure absolute synchronization between the scanning at the picture tube and that at the camera tube, the receiver contains circuits (called sync separator circuits) for separating the synchronizing pulses from the complete television signal. As is noted in the figure, these impulses are applied to the deflection circuits in the receiver and keep the two scanning beams—the one in the camera tube at the transmitting end and the other in the picture tube at the receiving end—in perfect synchronism. In this way the image of the scene is traced out by the moving spot of light on the screen of the picture tube.

A more detailed discussion of synchronizing circuits is given in chapter 8. This discussion is supplemented by examples of specific circuits. These examples are drawn from television receivers on the market today.

In the above description we have omitted a consideration of the sound broadcasting which almost invariably is a part of the television broadcast. For the present, it will be sufficient to understand that the sound is transmitted and received in the same way as a conventional sound broadcast even to the extent that an entirely separate carrier is used to carry the modulation of the sound accompanying the television broadcast.

THE TELEVISION SIGNAL

We are now in a position to consider more fully the nature of the signal which is used to transmit the television image. Up to this point we have explained in a general way that this signal contains the electrical image of the scene being televised and in addition contains the information required to synchronize the scanning at the camera tube with that at the picture tube. We will now go into greater detail as to the structure of this signal because of the bearing which it has on the operation and servicing of television receivers.

It will be helpful in understanding the nature of the television signal to review briefly the audio signal used in sound broadcasting and later to compare the two. In Fig. 1-18(A) is shown a typical sound or audio signal. As you know, the amplitude of this audio signal represents the intensity (loudness) of the sound, and the number of cycles per second represents the frequency (pitch) of the sound. During periods when the sound intensity is zero, the amplitude of the sound wave is, of course, zero; on the other hand, during periods when the sound intensity is high, there will then be a proportionate increase in the amplitude of the signal. An important characteristic which you should note is that in a sound wave, the amplitude of the wave has both positive and negative peaks and extends equally in both directions from the zero axis.

The Video Signal

A typical video or picture signal, that is, the electrical signal which represents the variations in brightness over the elements of a scene, is shown in Fig. 1-18(B). The horizontal line in the left-hand portion of the signal shown represents the signal voltage produced by an *unilluminated* or *black* portion of the scene, which is called the "black level." In discussing video signals this black level is used as a reference from which the light values corresponding to all other signal voltages are measured. The reason for this is apparent from inspection of the rest of Fig. 1-18(B), which represents various signal voltages corresponding to parts of the scene reflecting varying degrees of brightness. Note that as the brightness of the scene increases from black, the voltage of the video signal decreases, and that in any case the picture brightness never results in a signal voltage greater than the black level. The black level is therefore a suitable reference, because its voltage is fixed and easily reproduced. "White" would not be a suitable reference because the signal voltage corresponding to "whitest white" depends upon the maximum intensity of illumination available.

A video signal in which the signal voltage decreases (from the black level) as the picture brightness increases toward white is said to have a "negative picture polarity." The signal of Fig. 1-18 (B), as we have seen, is of this type. If we call the black-level voltage zero volt in Fig. 1-18 (B), all other shades of brightness will produce negative voltages, the whitest part of the scene having the greatest negative voltage. A signal having negative picture polarity is used to modulate the television carrier in present-day American television. Of course it is perfectly possible to have a video signal in which the signal voltage increases as the picture brightness increases; in this case the signal is said to have a "positive picture polarity." In fact we shall see later on, in considering receiver circuits, that the signal undergoes a reversal of polarity in passing through each stage of the video amplifier.

The important difference between the audio signal and the video signal is that the *video signal is always located on only one side of the black reference level,* whereas the audio signal contains variations on both sides of the zero-signal level. The video signal is therefore a pulsating voltage with a d-c component, not an a-c wave like the audio signal.

Let us now consider the video signal in greater detail. Fig. 1-19(A) shows the output of the camera tube for two successive lines of the image. At the same time Fig. 1-19(B) shows these two lines as they appear on the scanning pattern or raster. Starting at *a*, the beginning of the field, the beam traces the first line *a-c*; referring to the signal [Fig. 1-19 (A)] it can be seen that the image is black at

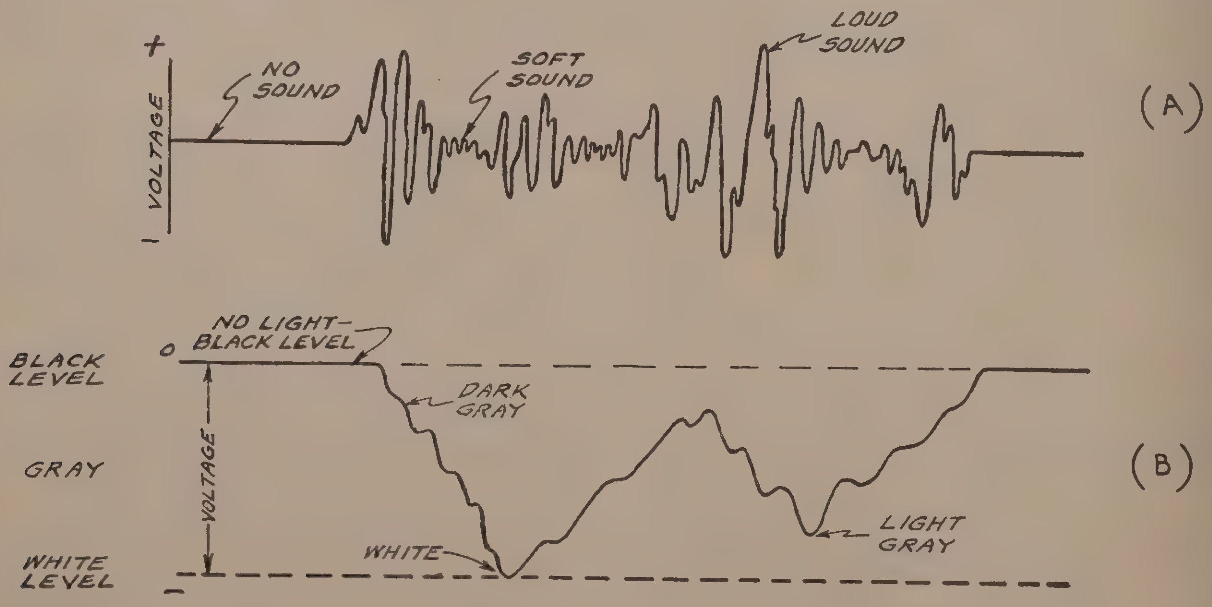


Fig. 1-18.—Comparison between a sound and video signal. In the video signal, black is represented by a fixed level, and various shades of brightness are represented by voltages displaced proportionately from the black reference level.

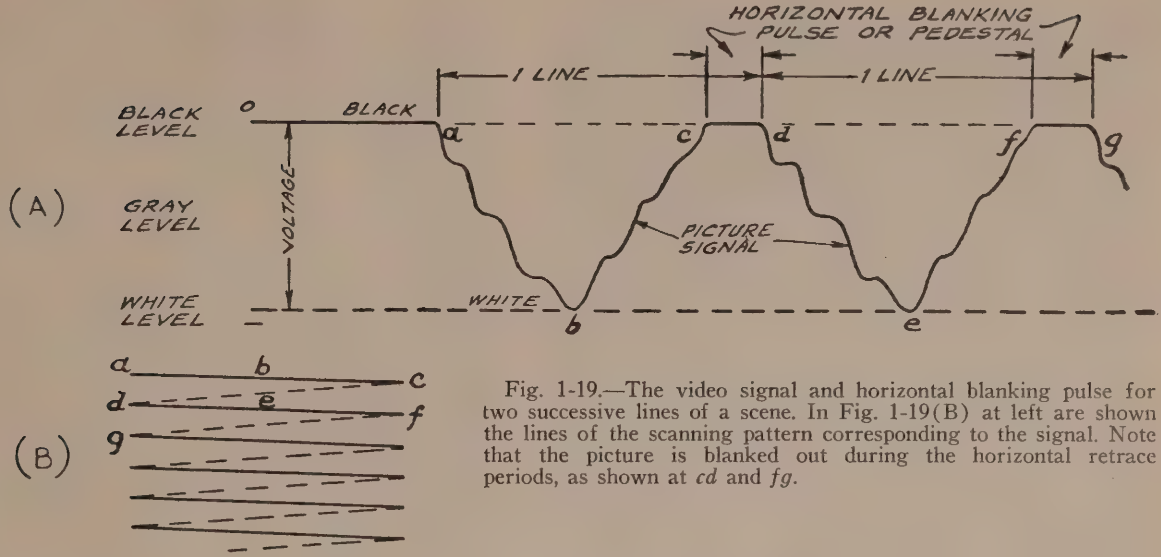

HORIZONTAL BLANKING PULSE OR PEDESTAL
1 LINE
1 LINE
BLACK LEVEL
O
BLACK
a
c
d
f
g
(A)
GRAY LEVEL
VOLTAGE
PICTURE SIGNAL
WHITE LEVEL
WHITE
b
e
(B)
a
b
c
d
e
f
g

Fig. 1-19.—The video signal and horizontal blanking pulse for two successive lines of a scene. In Fig. 1-19(B) at left are shown the lines of the scanning pattern corresponding to the signal. Note that the picture is blanked out during the horizontal retrace periods, as shown at *cd* and *fg*.

a, changes gradually to a brilliant white at *b*, and finally changes gradually to black at the end of the line *c*.

During the retrace time *c-d*, the signal level is maintained at a uniform black level as shown by *c-d* on Fig. 1-19(A) This "black" interval during which the retrace is carried out is called the "horizontal blanking period," and the pulse *c-d* is called the "horizontal blanking pulse," or the "horizontal blanking pedestal." As we shall see later in more detail, the pedestal performs two functions: (1) It blanks out the return trace so that it will not appear on the screen of the picture tube and (2) it provides a platform on which the horizontal synchronizing pulse is erected. Note that the second scanning line *d-e-f* is essentially the same as the first line and again the line is terminated on the signal wave by a blanking pulse *f-g*. In this way the entire field is scanned and the picture signal corresponding to the light and dark variations of the field is produced.

Signal and Sync Pulses

In the previous section we showed the video signal without discussing the modifications which must be made in the signal in order to provide the necessary synchronization. We shall now describe how this synchronizing information is added to the signal.

Consider the video wave shown in Fig. 1-20. This shows the wave for the last two lines of the field, just preceding the vertical retrace period during which the beam returns to the upper portion of the field. Starting at the left of the figure, the first thing you will note is the addition of a pulse which is erected on top of the horizontal blanking pulse; this is called the "horizontal sync pulse." As we shall see later in the discussion of receiver sync circuits, this small rectangular pulse provides the means which keeps the horizontal line or scanning oscillator in the receiver in synchronism with the scanning at the camera tube. An important thing

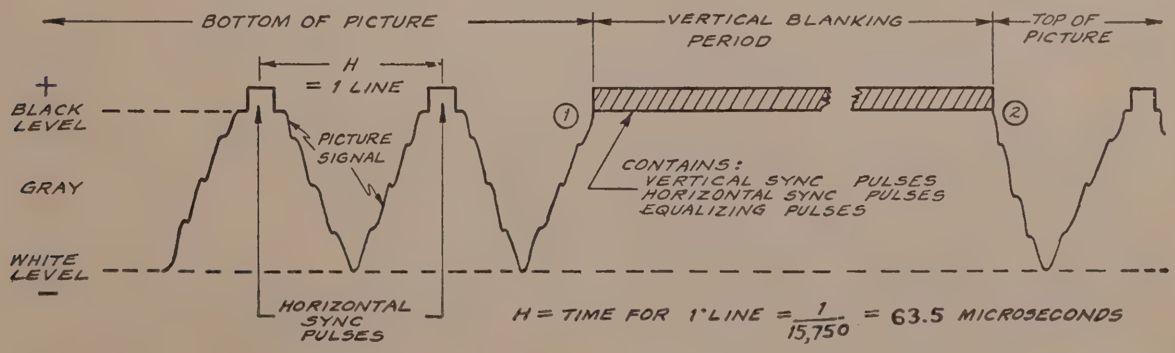

BOTTOM OF PICTURE
VERTICAL BLANKING PERIOD
TOP OF PICTURE
H = 1 LINE
+
BLACK LEVEL
PICTURE SIGNAL
①
②
CONTAINS: VERTICAL SYNC PULSES HORIZONTAL SYNC PULSES EQUALIZING PULSES
GRAY
WHITE LEVEL
−
HORIZONTAL SYNC PULSES
H = TIME FOR 1 LINE = 1/15,750 = 63.5 MICROSECONDS

Fig. 1-20.—The video signal showing the last two lines of a field followed by a vertical blanking period. Horizontal sync pulses are shown on the horizontal pedestals and the vertical sync pulse is transmitted during the vertical blanking period.

to note is that this sync pulse is located in the "blacker-than-black" region so that the screen of the picture tube is kept dark during the period of the horizontal retrace. Thus the portion of the signal more positive than black is used for synchronizing information, whereas the voltage more negative than the black level is used for the picture information. It is also important to observe that this line-synchronizing pulse appears at the end of each line so that constant synchronization of the line oscillator is maintained.

At the end of the last line at the bottom of the field, represented by point 1 in Fig. 1-20, the beam is ready to return to the upper edge of the field. As in the case of horizontal scanning, a sync pulse is required to return the beam to the top at the proper instant. All the information associated with the end of the field, required to return the beam to the top of the field in preparation for the one to follow, is contained in the interval designated as the "vertical blanking period." Essentially the following three functions are performed during the vertical blanking period: (1) A field synchronizing pulse is provided (the exact nature of this pulse is described later) so that the beam will be returned to the start of the frame at the proper instant. (2) The entire signal is blanked out so that the field retrace and lines scanned during this interval will not be apparent to the observer. (Actually, of course, the line-scanning circuits continue to function, but the beam does not exist because of the negative voltage on the control grid of the picture tube during this interval.) (3) The line synchronizing pulses are maintained during the vertical blanking period, which lasts for about 15 lines, so that the horizontal deflecting circuit in the receiver will not slip out of synchronism during this period. In addition to the vertical and horizontal sync pulses, two groups of so-called "equalizing pulses" are transmitted during the vertical blanking period; these are required for reasons which will be explained later.

Standard Television Signal

It is desirable at this point to show the complete signal which is used as the standard in this country. To illustrate the make-up of this signal, Fig. 1-21 shows the signal for two successive fields in the neighborhood of the vertical blanking pulse. Accordingly the left-hand portion of (A) shows the last four lines of any one field. This is followed by the vertical blanking period which contains equalizing pulses both preceding and following the vertical sync-pulse interval. After the last equalizing pulse, the horizontal sync pulses are resumed; by this time the beam has been returned to the upper portion of the screen so that shortly thereafter the normal video signal is resumed. To summarize, the first line (A) shows the complete signal as it exists for any one field both before and after the transmission of the vertical blanking period.

Part (B) of this figure describes the signal as it exists 1/60 second later for the following field. Since the scanning is interlaced, note that the line-sync pulses in (B) appear *between* the line-sync pulses in (A), thus providing the timing which is essential for interlacing. Again the last line in this field is followed by a vertical blanking period at the end of which the video signal is resumed. Note that for both parts of the figure, the reference point from which time is reckoned is the beginning of the vertical sync pulse, which is designated as taking place at any time represented by $t = t_1$. Using this time reference, it of course follows that the vertical sync pulse for the next field must begin 1/60 second later (since there are 60 fields per second); this is shown by part (B) of the figure so that the two vertical sync pulses are directly below each other but 1/60 second apart in time.

The description which follows shows in greater detail those parts of the complete video signal which have already been described.

Horizontal Blanking and Synchronization

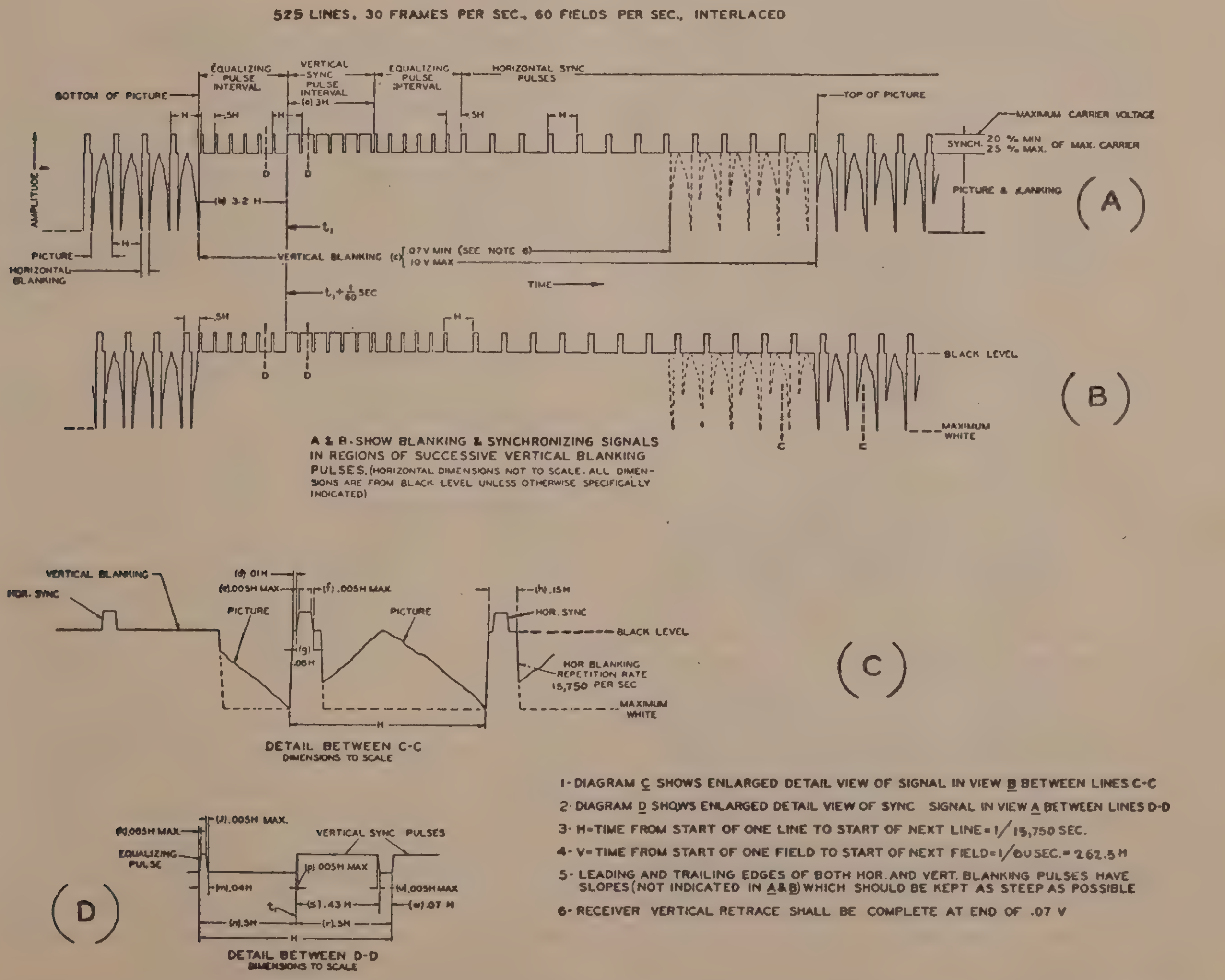
As shown in Fig. 1-21, the horizontal sync signal is transmitted at the end of each line and consists of an essentially rectangular pulse erected on the horizontal blanking pedestal. The amount of time allowed for the blanking pedestal is specified as 15% of the total time from the beginning of one line to the beginning of the next line. Since the time for each line (including the retrace) is 63.5 microseconds (1/15,750 second), the time devoted to horizontal blanking is about 9.5 microseconds. This interval of 9.5 microseconds has been found to be just large enough to allow for the retrace time, to allow the spot to assume normal scanning speed at the left edge of the picture, and to maintain reliable synchronization.

The whole of the horizontal blanking interval is not utilized for the synchronizing pulse as can be seen by examining the enlarged view of the wave between *C-C* [Part (B)] shown in the detail view, Part (C) of the figure. Actually only about half the total blanking time is used, and the front or leading edge of the sync signal is placed as close as possible to the beginning of the blanking pulse. The small allowance which is made takes care of some variation in timing and insures that the sync pulse will not run into the video portion of the signal and thus upset the line timing.

Vertical Blanking and Synchronization

The vertical blanking interval follows the last line of each field and consists of the following four parts which will be considered separately.

(1) Six equalizing pulses one-half line apart precede the sync pulse and accomplish (a) the maintenance of horizontal or line synchronization and (b) the "equalization" of the intervals preceding the vertical sync pulse so that *conditions preceding the vertical sync pulse are identical for alternate fields.* The need for these equalizing pulses arises because of the interlacing of alternate fields. As you can see from Fig. 1-21, the lines in the second field (B) are interlaced with those of the preceding field (A). If the equalizing pulses were eliminated and the vertical sync pulses were inserted in (A) at the end of the last line, then the vertical sync pulse would have to appear in the next field (B) at the middle of the line; the reason for this is that 1/60 second later the beam is in the middle of the line because of the interlacing. Thus *without the equalizing pulses,* the conditions preceding the vertical sync pulse would be different for each of the two fields. This would tend to produce a different type of vertical sync pulse for alternate fields, upset the synchronization and give rise to the distortion known as "pairing of the interlace."


525 LINES, 30 FRAMES PER SEC., 60 FIELDS PER SEC., INTERLACED
EQUALIZING PULSE INTERVAL
VERTICAL SYNC PULSE INTERVAL
(o) 3H
EQUALIZING PULSE INTERVAL
HORIZONTAL SYNC PULSES
BOTTOM OF PICTURE
TOP OF PICTURE
MAXIMUM CARRIER VOLTAGE
SYNCH. 20 % MIN 25 % MAX. OF MAX. CARRIER
H
.5H
H
.5H
H
D
D
AMPLITUDE
(b) 3.2 H
t_1
PICTURE & BLANKING
(A)
PICTURE
HORIZONTAL BLANKING
H
VERTICAL BLANKING (c) .07V MIN (SEE NOTE 6) .10 V MAX
TIME
$t_1 + \frac{1}{60}$ SEC
.5H
H
D
D
BLACK LEVEL
(B)
MAXIMUM WHITE
C
C
A & B - SHOW BLANKING & SYNCHRONIZING SIGNALS IN REGIONS OF SUCCESSIVE VERTICAL BLANKING PULSES. (HORIZONTAL DIMENSIONS NOT TO SCALE. ALL DIMENSIONS ARE FROM BLACK LEVEL UNLESS OTHERWISE SPECIFICALLY INDICATED)
VERTICAL BLANKING
HOR. SYNC
(d) .01H
(e) .005H MAX
(f) .005H MAX
PICTURE
PICTURE
(g) .08H
(h) .15H
HOR. SYNC
BLACK LEVEL
HOR BLANKING REPETITION RATE 15,750 PER SEC
MAXIMUM WHITE
H
(C)
DETAIL BETWEEN C-C
DIMENSIONS TO SCALE
(k) .005H MAX
(j) .005H MAX.
EQUALIZING PULSE
VERTICAL SYNC PULSES
(p) .005H MAX
(m) .04H
(u) .005H MAX
t_1
(s) .43 H
(v) .07 H
(n) .5H
(r) .5H
H
(D)
DETAIL BETWEEN D-D
DIMENSIONS TO SCALE

1- DIAGRAM C SHOWS ENLARGED DETAIL VIEW OF SIGNAL IN VIEW B BETWEEN LINES C-C
2- DIAGRAM D SHOWS ENLARGED DETAIL VIEW OF SYNC SIGNAL IN VIEW A BETWEEN LINES D-D
3- H=TIME FROM START OF ONE LINE TO START OF NEXT LINE=1/15,750 SEC.
4- V=TIME FROM START OF ONE FIELD TO START OF NEXT FIELD=1/60 SEC.=262.5 H
5- LEADING AND TRAILING EDGES OF BOTH HOR. AND VERT. BLANKING PULSES HAVE SLOPES (NOT INDICATED IN A&B) WHICH SHOULD BE KEPT AS STEEP AS POSSIBLE
6- RECEIVER VERTICAL RETRACE SHALL BE COMPLETE AT END OF .07 V

Fig. 1-21—The standard television signal. Part (A) shows the signal at the end of any one field; part (B) shows the signal at the end of the next field, 1/60 second later. The difference between the two fields is caused by interlacing.

In a paired interlace, the even-scanned lines do not lie midway between the odd-scanned lines, because of the difference in timing on alternate fields.

In connection with the maintenance of line synchronization during the vertical blanking interval, note that the leading edges of the equalizing pulses function to maintain synchronization. Not all the pulses are used for each field, however. Thus note that because of the interlacing, the first, third, and fifth equalizing pulses are used on the first field (A), and the second, fourth, and sixth pulses are used on the succeeding field (B). This explains why six pulses are used, each spaced one-half line apart, rather than three pulses spaced one line apart. It would, of course, be possible to use three different pulses in each field, but if this were done the signal preceding the vertical sync pulse would be different for succeeding fields and there would be a resulting absence of equalization.

(2) The vertical sync pulse (or field synchronizing pulse) follows directly after the equalizing pulse interval and consists of six broad pulses in which the edges are serrated or cut at one-half line intervals. The function of the vertical sync pulse is to provide the control signal which tells the vertical oscillator that it is time to begin the retrace and thus to return the beam to the top for the beginning of the next field. The pulses in the vertical sync-pulse interval are considerably broader than the line pulses so that the sync separator circuit will be able to distinguish between the two types of pulses and thus be able to separate the vertical sync pulses from the horizontal sync pulses. At the same time the edges of the serrations at half-line intervals provide the necessary control for maintenance of horizontal synchronization. The serrations are required at half-line intervals because of the interlacing; the reasoning used in connection with the equalizing pulses also applies here.

(3) It was explained previously that in order to provide identical conditions for the two successive fields preceding the vertical sync pulses, six equalizing pulses were inserted in front of the vertical sync pulse in each field. It is just as necessary to keep the conditions following the vertical sync-pulse interval the same for the two successive fields (A) and (B); for this reason six lagging equalizing pulses appear in both (A) and (B) after the vertical sync pulse. If you examine the vertical sync-pulse interval in both (A) and (B) you will see that, although the lines in the two fields are displaced by one-half line because of the interlacing, nevertheless the conditions in the neighborhood of the vertical sync-pulse interval are the same for both fields. Note that the lagging equalizing pulses are also one-half line apart so that line synchronization is maintained for both the "odd" and "even" fields.

(4) The lagging equalizing pulse interval is terminated before the end of the vertical blanking period so as to prepare the line oscillator for the normal horizontal sync pulses which are to follow. In practice, the video signal is blanked out for a period of from 7 to 12 lines following the last equalizing pulse so that the line oscillator (which may have been operating at double line frequency during the preceding period) has a chance to settle down to being under control of the normal type of sync signal. At the end of the vertical blanking interval, the blanking is of course removed and the video portion of the signal again controls the intensity of the beam in the picture tube.

Range of Frequencies in Video Signal

Unlike audio signals which contain frequency components ranging from a low value of about 20 cycles per second to a high value of about 15,000 cycles per second, video signals include a range from practically zero frequency (produced over areas where there is little variation in light intensity) to as high as 4 or more megacycles (produced over areas where there is a very rapid variation in light intensity).

It is interesting to examine the manner in which the maximum frequency required in the video signal is related to the amount of detail which is reproduced. This can be arrived at from the following considerations: Let us assume that we wish to transmit a picture in which the same resolution or detail is desired in the horizontal direction as in the vertical direction. In the vertical direction we have a total of 525 lines or 525 elements. Since a horizontal line is 4/3 times as long as a vertical line (the picture is 4/3 times as wide as it is long), it follows that there are 4/3 as many elements in a horizontal line as there are in a vertical line. This makes a total of 525 x 525 x 4/3 or 367,500 elements in the complete picture.

Since we wish to calculate the maximum frequency which is required to reproduce the light and dark variations over each one of these 367,500 elements, let us assume that alternate elements in the picture are black and white so that the image resembles a checkerboard pattern. This type of scene requires the highest possible frequency for faithful reproduction because of the rapid variation in light intensity as the beam goes from one element to the next. The exact opposite of this type of scene would be one which was uniform over its entire area, for in this case it would be necessary for the system to transmit only very low frequencies. For the checkerboard pattern under discussion, the variation from black to white in scanning two adjacent elements requires a certain amount of time. This amount of time represents the duration of one complete cycle and one divided by this number represents the highest frequency which must be transmitted.

Let us calculate this time interval required for the beam to scan two adjacent black and white elements. Since there are 367,500 elements in a complete picture and it requires 1/30 second to transmit this picture, the time allotted to the transmission of information on two elements (1 cycle) is equal to

$$\frac{2\sqrt{2}}{367{,}500} \times \frac{1 \text{ second}}{30 \text{ cycles}} = \frac{1 \qquad \text{second}}{3{,}900{,}000 \quad \text{cycles}}$$

The constant, $\sqrt{2}$, is determined by several factors, the most important of these is the spot diameter. The frequency generated when the scanning beam passes over these two elements is thus equal to 3,900,000 cycles per second, or 3.9 megacycles.

As we have previously seen, the maximum frequency which is present in the video signal is related directly to the amount of detail required in the image. It has been found that for the smaller picture tubes satisfactory detail is obtained when frequency variations up to 2.5 megacycles are transmitted and that little is gained by transmitting the high frequency components produced in the scanning at the camera tube. However, where a comparatively large picture tube is used, additional detail and a finer image can be produced by transmitting frequency components ranging up to about 4 megacycles.

The Modulated Wave

In previous sections we have described the video wave produced when a scene is scanned and the modifications which are made in this wave in order to provide for both line and field synchronization. We now will consider the make-up of the modulated wave which is produced when this video wave modulates a high-frequency carrier.

First yet us review the result of amplitude modulating a carrier with a conventional audio signal. As Fig. 1-22(A) shows, if we amplitude modulate a 1000-kc carrier with an audio signal which contains frequency components ranging up to 5 kc, then the resulting modulated wave contains, in addition to the carrier frequency, new frequencies which extend to the limits of 5 kc below the carrier frequency and 5 kc above it. In other words, the process of amplitude modulating the carrier results in the introduction of two sets of *sidebands* which extend outward from the carrier to a value equal to the highest frequency in the modulating wave. The sideband which contains the frequencies lower than the carrier frequency is called the "lower sideband," and that which contains the frequencies higher than the carrier is called the "upper sideband."

It would be impossible to use a broadcast-band carrier for television work. This can be seen from the fact that unlike sound, the video modulating frequencies would themselves be higher than the carrier frequency. In order to make the modulation process work, it is necessary that the carrier frequency be at least several times the highest frequency of modulation. For this reason the carrier frequency in television must be several times as high as the maximum video frequency, or several times 4 mc. Actually the carrier frequencies which have been chosen for television are at least ten times the highest video frequency since the lowest carrier frequency which is being used for 525-line television is 44 mc.

Some of you may recall that a number of years ago, carrier frequencies only a little higher than the

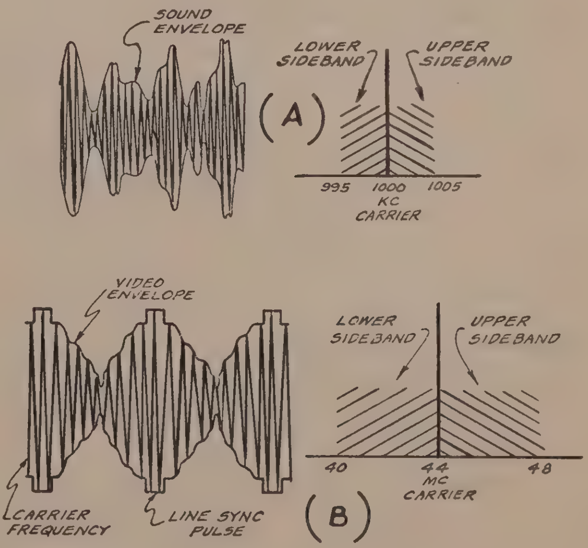


Fig. 1-22.—The sidebands in an a-m sound signal occupy a bandwidth of only about 10 kc, whereas the sidebands in a television signal require a bandwidth of about 8 mc.

broadcast-band were assigned to experimental television. This merely illustrates our point, however, since those early low-definition pictures contained comparatively few elements and consequently had maximum video frequencies far below those found in present-day high-definition work.

Returning to the comparison between a broadcast-band carrier amplitude modulated by a sound wave and a modulated television carrier, we show in Fig. 1-22(B) the wave which results when a 44-mc carrier is modulated with a video signal. As in the case of the 1000-kc carrier, the process of modulation introduces two sets of sidebands, and for the example shown the two sidebands extend to 4 mc below and 4 mc above the 44-mc carrier.

It is interesting to compare the bandwidth required for the transmission of a scene by television with the bandwidth required in amplitude modulated (a-m) sound broadcasting. As Fig. 1-22 shows, an a-m sound broadcast requires only a 10-kc channel whereas the television channel (arranged for double-sideband modulation) requires 8000 kc or 800 times as much space as the sound channel. Although it is possible to locate approximately 100 sound channels in the broadcast band, it would require more than 8 times the space provided by the entire broadcast band for the transmission of a single television channel with double-sideband modulation. The large amount of the radio spectrum required for television is one of the reasons for the choice of the ultrahigh frequency range for television.

Positive and Negative Modulation

In discussing video signals we pointed out that the video signal is said to have a positive or negative picture polarity depending upon whether the changes from the black level take place in a positive or a negative direction, respectively. In modulation we run into somewhat similar terms—positive and negative modulation—which are related not to the polarity of the video signal but to the modulated wave itself. A television carrier is said to have positive modulation when an increase in carrier amplitude corresponds to a brighter area in the scene being scanned. Thus for a wave with *positive* modulation, the *lowest* carrier amplitude corresponds to *black* while the maximum carrier amplitude corresponds to the brightest part of the image. On the other hand, for negative modulation, a decrease in carrier amplitude corresponds to an increase in the brightness of the image. Thus for a wave with *negative* modulation, the lowest carrier amplitude corresponds to maximum white in the image and the *highest* carrier amplitude corresponds to *black*.

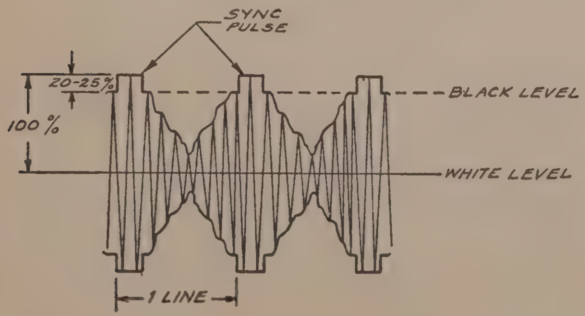


Fig. 1-23.—A video wave with negative modulation. The sync pulses are located in the blacker-than-black region and occupy from 20 to 25 percent of the maximum carrier amplitude.

Because negative modulation offers certain advantages in improved performance and simplified receiver design, it is used as standard in this country. As shown in Fig. 1-23, the maximum amplitude of the carrier is used for the synchronizing pulses and lies in the blacker-than-black region. The term "blacker-than-black" merely means that the sync signals have higher amplitude than black picture signals. It has been found that reliable synchronization can be secured so long as the amplitude of the synchronizing pulses is from 20 to 25 percent of the maximum carrier amplitude. Actually, then, not more than 80 percent of the total carrier amplitude is available for transmitting information on the light values in the scene, the rest of the wave being used for synchronization.

RECEIVER CIRCUITS: GENERAL

Having examined the fundamental principles of television, let us now investigate the operation of receiver circuits. These circuits are especially important because the work of servicemen in the field deals primarily with the installation and maintenance of receivers. In order to show the interrelationship between the many components that make up a receiver, we shall first break down the receiver into its major sections and later consider the functioning of these in more detail.

Fig. 1-24 shows a block diagram of a typical television receiver arranged to show the general character of the signal and the function performed by each section. For convenience we shall assume that the receiver is tuned to the 44-50 mc channel. In accordance with the preceding description, this means that the frequency of the video carrier is 45.25 mc (1.25 mc above the low-frequency end of the channel) whereas the frequency of the audio carrier is 49.75 mc (0.25 mc below the high-frequency end of the channel).

Both these signals, together with their sidebands, are picked up by the antenna and fed through a transmission line to the input of the r-f amplifier. Essentially the function of the r-f amplifier is the same as that of the r-f amplifier in any superheterodyne receiver—to amplify the signal and to reject unwanted signals in adjacent and other channels. In this case, the r-f amplifier is broadly tuned so that both the video and sound carriers, which are separated by 4.5 mc, are amplified equally.

After being amplified in the r-f amplifier, both signals are fed to the mixer circuit where the conversion of the signals to the intermediate frequencies (i.f.) takes place. Since there are two radio frequencies, it of course follows that two separate intermediate frequencies are produced.

In accordance with present practice, the oscillator operates at a frequency approximately 26 mc *above the video carrier frequency.* For the channel being received, the frequency of the oscillator in the receiver is 71 mc. Since the oscillator frequency is 25.75 mc above the video carrier frequency, in this case, 25.75 mc will be the value of the video i.f. In the same way, the i.f. of the sound signal is equal to the difference between the oscillator frequency and the sound carrier frequency, 71—49.75 mc, or 21.25 mc.

Following the mixer, the sound channel in the block diagram is entirely independent of the rest of the receiver and in practically every detail is similar to a conventional receiver used for broadcast reception. Thus the 21.25-mc sound i-f signal passes through the sound i-f amplifier (the selectivity of which is broader than usual to minimize the effects of oscillator drift), and is demodulated at the sound detector. The avc voltage is supplied in the usual manner to control the gain of the stages in the sound i-f amplifier. In accordance with FCC standards, frequency modulation (f.m.) is used exclusively for all television sound.

Returning to the video signal, we have seen that a 25.75-mc i-f signal is produced by the mixer and that this signal carries the video modulation. As the diagram shows, this signal is amplified in the *video i-f amplifier,* which usually consists of several stages, and finally reaches the *video detector* where the signal is demodulated. The video signal recovered at this point is essentially the same as the output of the camera tube, so that it contains all the information required to reproduce the picture, and in addition, includes the blanking and sync pulses. The video detector is followed by the *video amplifier* which, in terms of a sound receiver, corresponds to the audio amplifier. The function of the video amplifier is to amplify the video signal so that its amplitude will be great enough to "swing" the modulation grid of the picture tube. For the average picture tube this requires approximately 25 volts, peak-to-peak.

Note in the diagram that the polarity of the video signal is reversed 180 degrees for a single stage of video amplification and that the receiver is arranged so that the signal which reaches the control grid of the picture tube has a positive polarity. As a result the synchronizing impulses appear in the blacker-than-black (highly negative grid-bias) part of the picture-tube characteristic so that the beam is blocked during the retrace part of the line and field sweeps.

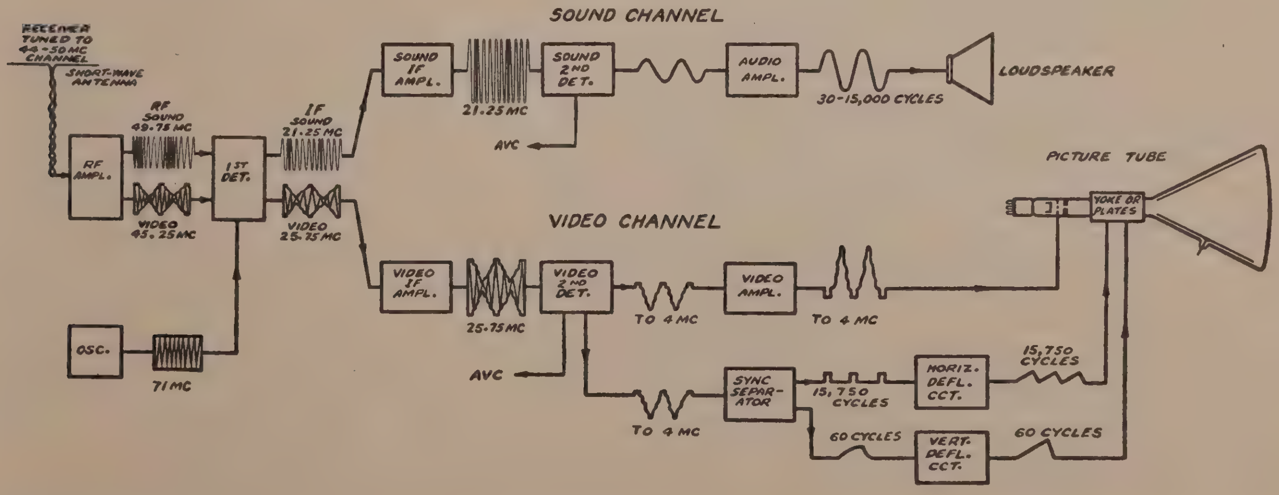

SOUND CHANNEL
RECEIVER TUNED TO 44-50 MC CHANNEL
SHORT-WAVE ANTENNA
RF SOUND 49.75 MC
RF AMPL.
VIDEO 45.25 MC
1ST DET.
IF SOUND 21.25 MC
VIDEO 25.75 MC
OSC.
71 MC
SOUND IF AMPL.
21.25 MC
SOUND 2ND DET.
AVC
AUDIO AMPL.
30-15,000 CYCLES
LOUDSPEAKER
VIDEO CHANNEL
VIDEO IF AMPL.
25.75 MC
VIDEO 2ND DET.
TO 4 MC
AVC
VIDEO AMPL.
TO 4 MC
PICTURE TUBE
YOKE OR PLATES
TO 4 MC
SYNC SEPAR-ATOR
15,750 CYCLES
HORIZ. DEFL. CCT.
15,750 CYCLES
60 CYCLES
VERT. DEFL. CCT.
60 CYCLES

Fig. 1-24.—A block diagram of a typical television receiver. Note the changes in the signal as it passes through the receiver.

In addition to supplying the video signal and the signal which actuates the avc system, the detector supplies the video signal to the *synchronizing separator*. The purpose of this separator is to remove the picture component from the complete video signal and then to separate the horizontal sync pulses from the vertical sync pulses. As is shown, the horizontal sync pulses are arranged to control the timing of the horizontal deflection circuit, while the vertical sync pulses are arranged to control the timing of the vertical deflection circuit.

The power supply is not shown in the block diagram. A single low-voltage power supply can be used to take care of all voltage requirements throughout the receiver with the exception of the high-voltage requirements for the picture tube. The latter, which may include voltages as high as 30,000 volts, is supplied by a separate high-voltage power supply which has its own transformer, rectifier, and filter.

The two most popular high-voltage power supplies in use at the present time are the "kick-back" power supply and the r-f power supply. Television power supplies are covered in chapter 11.

And now, having discussed the fundamental principles of television here, detailed coverage of the various television receiver circuits will be given in succeeding chapters.

CHAPTER 2

FREQUENCY CHARACTERISTICS OF THE TELEVISION SYSTEM

BY SEYMOUR D. USLAN

In order to understand many of the features of television it is very important to know something about the frequency standards concerning television broadcasting as set up by the Federal Commuications Commission (FCC). Such information will be very helpful because in the discussions to follow reference is quite often made to these standards. The frequency range of television, the different channel frequencies, the bandwidth involved, and the video and sound carrier frequencies are some of the topics which will be discussed in this chapter.

Television Channels

Television today occupies quite a large frequency range. The FCC has set aside 13 *television channels* in the frequency spectrum between 44 and 216 megacycles. These 13 television channels do not exist over this complete frequency range but only over part of it. The term "television channel" according to the FCC means—a band of frequencies 6 megacycles wide in the television broadcast band and designated either by number or by the extreme lower and upper frequencies. This means then for the 13 television channels, the television broadcast band occupies 13 × 6 or 78 mc. However, this 78-mc band is not continuous as are the f-m and a-m commercial broadcast bands.

From the 44-to-216-mc frequency range the following frequencies are not part of the television band: 50-54 mc, 72-76 mc, and 88 to 174 mc. The other frequencies in this 44-to-216-mc frequency spectrum are assigned to the television band. These remaining frequencies are divided into 13 television channels and each channel is designated by a number. In other words the channels are numbered 1 through 13. Consequently with the width of each television channel standardized at 6 mc, the frequency ranges of the television channels are shown in the accompanying table.

CHANNEL NO.	FREQUENCY (MC)	CHANNEL NO.	FREQUENCY (MC)
1	44 — 50	7	174 — 180
2	54 — 60	8	180 — 186
3	60 — 66	9	186 — 192
4	66 — 72	10	192 — 198
5	76 — 82	11	198 — 204
6	82 — 88	12	204 — 210
		13	210 — 216

These television channels are often divided into two separate groups. Channels 1 through 6 are considered as one group, often termed the low band, and channels 7 through 13 are designated as the other group, usually called the high band. Such a grouping is readily understandable when one considers the 86-mc difference between the end of the sixth channel and the beginning of the seventh. Such a grouping is helpful in analyzing image frequency interference and other problems.

Video Signal Characteristics

When a (high frequency) carrier is modulated with a video signal, the carrier undergoes changes in amplitude. Thus we see that television is another form of amplitude modulation and possesses many of the characteristics of regular a-m radio broadcasting. In this latter type of broadcasting, when a carrier signal is modulated by audio the resultant modulated signal consists of a center frequency component, equal in frequency to the carrier, and two sideband frequency components. One sideband component is called the upper sideband and the other is called the lower sideband. The frequencies of these sidebands are equal to the center frequency plus or minus the frequency of the applied audio modulating signal. The relative amplitudes of the components of the a-m signal are determined by the percentage of modulation. The frequency separation between the components is determined by the frequency of the audio signal. In a-m radio broadcasting (550 to 1500 kc) the adjacent channel separation is 10 kc as set down by the FCC. This means that there is a maximum of only 5 kc on either side of an assigned station frequency.

In regular a-m broadcasting it is possible to use audio modulating frequencies as high as 15 kc. The frequency of the audio modulating signal determines the bandwidth of the transmitted signal.

However, since the maximum bandwidth as set down by the FCC is equal to 10 kc, then the maximum audio frequency that can be passed is 5 kc.

With this much understood we can illustrate the frequency spectrum of a typical a-m wave (audio modulated) with regard to both amplitude and frequency. In Fig. 2-1 we have illustrated the frequency spectrum of a 100-percent-modulated a-m wave. In a 100-percent a-m wave, the amplitudes of the upper and lower sidebands each is equal to half that of the carrier component. In Fig 2-1 the frequency of the lower is designated as f_L and that of the upper sideband f_U, and for maximum audio transmission they are shown as being 5 kc on either side of the center frequency component which is designated as f_o. If the frequency of the carrier is 1000 kc, then for a 5-kc audio modulating signal the upper and lower sideband frequencies are respectively 1005 kc and 995 kc. Thus an over-all bandwidth of 10 kc exists.

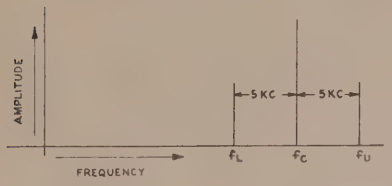


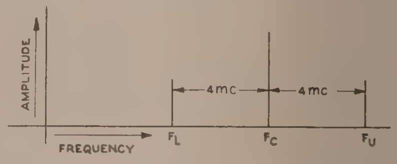
Fig. 2-1. — The frequency spectrum of a typical 100-percent-modulated a-m wave. The amplitude of the sidebands is half that of the carrier.

The intelligence (i.e. audio signal) is completely vested in the sideband components, with each sideband containing half of the power of the transmitted intelligence. (The center frequency component contains no intelligence at all.) Upon demodulation of the a-m signal, each sideband contributes to the amount of audio power output. If only one sideband were detected, all the audio frequencies involved at the transmitter would be reproduced but only half the audio power output would be obtained as compared with that resulting when both sidebands are detected. Consequently we can see that whether one or both of the sidebands are detected all the audio frequencies will be reproduced and only the power output at these frequencies will be different.

In video signal transmission the situation is very much the same regarding the intelligence and the sidebands. In television transmission, the picture signal (i.e. video signal) is similar to the audio signal in a-m transmission. The video signal is, therefore, used to amplitude modulate a high-frequency carrier signal. The resultant modulated signal consists of a center frequency component and two sidebands, an upper and a lower. The frequency of the sidebands will be equal to the frequency of the carrier plus or minus the frequency of the video signal. All the video intelligence is vested in the sidebands and none in the center frequency component. Conditions for the percentage of modulation in television are the same as in regular a-m broadcasting.

The video frequencies are as low as 30 cycles and as high as 4 mc. This represents quite a problem for the design of proper circuits to pass such a band of frequencies. The frequency spectrum picture of the video-modulated television signal for 100-percent modulation is illustrated in Fig. 2-2. In this drawing F_C is the carrier frequency, F_L the lower sideband, and F_U the upper sideband. The maximum bandwidth is shown to be equal to 8 mc. This video bandwidth plus that of the sound f-m signals would take up too much space in the assigned frequency spectrum for television. As mentioned, according to the FCC, each television channel occupies an over-all bandwith of 6 mc. This includes the video-modulated signals and the sound f-m signals.

Fig. 2-2. — The frequency spectrum of a television signal at 100-percent modulation. Note that each sideband is spaced 4 mc from the carrier.



Vestigal-Sideband Transmission

Since the maximum 8-mc bandwidth of the video-modulated signal is in itself too broad for the assigned television channels, some form of adjustment must be made. We have previously stated that if only one sideband of an audio a-m signal were transmitted, the primary difference in the receiver output would be in the amount of power. In other words all the audio modulating frequencies would be reproduced but with half the power they would have under normal transmission (double-sideband transmission).

Such a system of single-sideband transmission is also possible in television. The maximum bandwidth necessary for the video-modulated signal would then be only 4 mc—half that of the 8-mc bandwidth under double-sideband transmission. Single-sideband transmission is not used in television because it may cause distortion in the output of the detector system of the receiver and also because it is difficult to obtain the necessary sharp cutoff filter for complete suppression of one sideband. Since the television channels are 6 mc wide and since one sideband occupies a width of only 4 mc, approximately 2 mc (discounting the small bandwidth required for the sound f-m signal) will not be used if single-sideband transmission is employed. By utilizing the the extra bandwidth we can transmit a television signal that has one complete sideband plus part of the other. Such transmission is often called vestigal-sideband transmission, and it is the type employed in television today.

Vestigal-sideband transmission (sometimes called sesqui-sideband or quasi-single-sideband) refers to an a-m wave where one sideband is partially suppressed and the other sideband not suppressed at all. The suppression of the one sideband is such that the transmitted video-modulated signal in conjunction with the sound f-m signal occupies a bandwidth that just conforms to the 6-mc television channel bandwidth.

To understand the character of the transmitted video-modulated signal as far as its bandwidth is concerned, let us refer to an ideal television channel response as seen in Fig. 2-3. This response curve is assumed to be that for channel number 1, although it will hold for all the other channels with the proper upper and lower frequency limits. This curve represents the ideal characteristic of the transmitted television signal as well as that of the r-f section of the television receiver. According to the FCC, the sound carrier shall have a frequency 0.25 mc lower than the upper frequency of the channel in question. This means that for channel number 1 (44-50 mc) the sound carrier frequency would be 49.75 mc. The FCC also stipulates that the video carrier frequency shall be 4.5 mc lower in frequency than the sound carrier frequency. This means that for channel number 1 the video carrier would be at a frequency of 45.25 mc. Since each channel is 6 mc wide and we have already used 4.75 mc of it (from 45.25 to 50 mc) all that remains is 1.25 mc, which is at the low frequency end of the channel. For channel number 1 this 1.25-mc limit is between 44 and 45.25 mc.

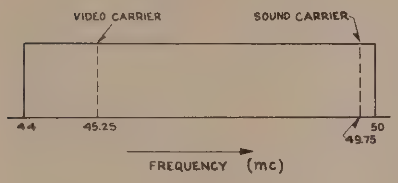


Fig. 2-3.—Ideal response curve for television channel 1. The curve is 6 mc wide, with the video carrier 1.25 mc away from the lower end of the band and the sound carrier 4.5 mc above the video carrier.

Since a double-sideband video-modulated signal will have a total bandwidth of 8 mc with each sideband having video frequencies up to 4 mc, then from the ideal response characteristic of Fig. 2-3 we see that the lower frequency sideband is the one that is partially suppressed. The upper frequency sideband will be at a frequency limit of 4 mc plus the video carrier signal and for channel number 1 this will be 4 plus 45.25 or 49.25 mc. Thus, from Fig. 2-3, this upper sideband is seen to fall well within the frequency response of the channel without infringing upon that part reserved for the sound f-m signal which will be at a maximum of 50 kc wide. Consequently we see that in the vestigal-sideband transmission of a video-modulated signal, it is the lower sideband which is partially suppressed without suppression of the upper sideband.

Examining the video part of the frequency response characteristic of Fig. 2-3 brings a few pertinent facts to light. It is the high-frequency end of the lower sideband that is suppressed and the low-frequency end that remains. (Those parts of the sidebands closest to the video carrier signal contain the low video frequencies. This means that in this type of vestigal-sideband transmission the higher frequency video signal components are weakened, which effectively means that the lower frequency components are strengthened. Much of the video energy is located primarily in the low-frequency signals. These low-frequency components of the video signal determine the background of the receiver picture detail.

Operating Bandwidth Characteristics

The bandwidth characteristics of Fig. 2-3 are ideal and in practice the straight line sides of the response curve are not obtained. In the television transmitter, special filter networks are employed to suppress the lower frequency sideband and the side of the response curve that results is not a vertical line as was shown in Fig. 2-3. It is quite difficult and expensive to design a filter that has a straight vertical line cutoff characteristic. For single-sideband transmission without any frequency discrimination of the sideband passed and likewise without passing any part of the suppressed sideband, a very sharp cutoff filter would have to be employed. With vestigal-sideband transmission in television, such a sharp cutoff filter is no longer necessary.

In the television system we have to deal with a number of frequency response curves to understand fully the different bandwidth characteristics of the system. There is the over-all video-modulated response of the video or so-called television transmitter, the response of the sound f-m transmitter, the response of the input or r-f section of the receiver, and the over-all video response of the receiver. This last frequency response characteristic includes that of the video i-f stages as well as the r-f stage. For most practical purposes the response of the r-f section of the receiver should be a combination of the video-modulated and sound f-m transmitted signals.

The sound f-m signal itself occupies a very small part of the 6-mc channel. Its maximum bandwidth is 50 kc (25 kc on either side of its center frequency) for 100-percent modulation. Let us refer to Fig. 2-4 which is an over-all picture of the bandwidth of a 6-mc channel which takes into account the video-modulated and sound f-m signals. This bandwidth characteristic is illustrated for 0 to 6 mc so it will be applicable to any channel. To use it for any channel all that must be done is to apply the proper frequency limits. The left-hand side of the curve illustrates the video response and the right-hand side shows that of the sound f-m response. The video response is seen to have sides that are sloping rather than straight lines. The upper video frequency sideband which is a maximum of 4 mc wide lies in the channel between 1.25 and 5.25 mc. The sound f-m signal has a maximum bandwidth of 50 kc (25 kc on either side of its center frequency) and thus occupies only a small portion of the channel.

The 0.5-mc separation between the sound carrier and the upper limit of the video frequency sideband is enough separation to prevent any overlapping of the video and sound f-m signals. This is better illustrated in Fig. 2-5 which is an enlargement of the upper frequency end of the television channel. The frequency separation between the 5.25-mc point and the 5.725-mc lower frequency limit of the f-m signal is 0.475 mc. This separation is sufficient to prevent any interaction between the limits of the video and sound signals. The distance between the 5.775-mc upper frequency limit of the f-m signal and the 6-mc limit of the television channel is 0.225 mc. This separation is often referred to as the *guard band* of the television channel. This guard band is intended to prevent any interaction between adjacent television channels.

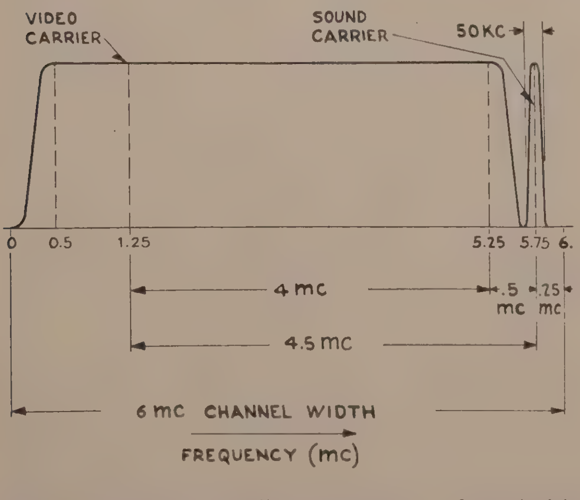


Fig. 2-4.—Actual over-all response curve of a television channel. Note that the response drops off and then rises again to a sharp peak at the sound carrier.

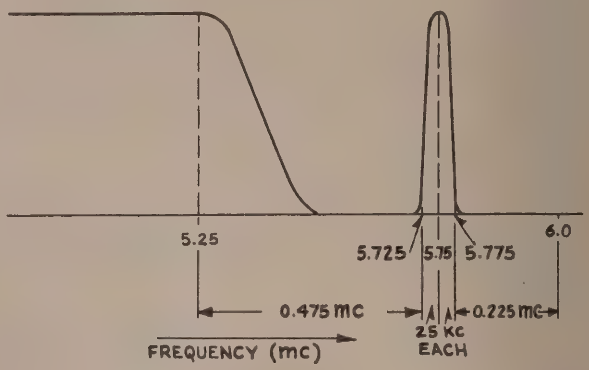


Fig. 2-5.—An enlargement of the upper frequency end of the television channel.

The video modulation frequencies should be so reproduced in the television receiver that they have the same relative amplitudes as when modulating the carrier at the transmitter. This is important so that the reproduced picture at the receiver takes on the same appearance as the scene at the studio. In vestigal-sideband transmission, however, the low video frequencies are transmitted double sideband, and upon reproduction these low video frequencies will consequently be stronger than the high video frequencies. In order that the video frequencies of the received signal be a faithful reproduction of those at the studio, the receiver incorporates within its video channel correction circuits in which the low video modulating frequencies are corrected for proper reproduction.

The low video frequencies must undergo a certain amount of attenuation so that the signal being applied to the picture tube will be a reproduction of that at the output of the camera tube. Fig. 2-6 illustrates the over-all bandwidth characteristics of the video channel of the television receiver. That part of the curve to the right of point *A* is the same as that portion of the video section of the curve of Fig. 2-4. Point *A* falls about 2.5 mc away from the low-frequency end of the channel. Between zero and this 2.5-mc point, the characteristic of the curve of Fig. 2-6 differs appreciably from that of Fig. 2-4. This new curve is such that at the video carrier frequency it will be down 50 percent or 6 db. In order that there be faithful reproduction of all the frequencies, the shape of the curve is such that for all frequencies there will effectively be 100-percent transmission. The reference to 100-percent transmission means the average level of the response characteristic, such as that at the high video frequency end of the band.

For instance, the low video frequencies of the upper sideband are attenuated in a gradually decreasing manner to about 1.25 mc from the carrier frequency. This means that as the video frequencies from 0 to 1.25 mc of the upper sideband are increased, the amount of attenuation is decreased. To have the reproduction of these low video frequencies at so-called 100-percent transmission throughout the entire channel, those low video frequencies on the suppressed or lower sideband have to compensate for the attenuation of those of the upper sideband. This compensation must, however, be in proportion to the amount of attenuation of the upper sideband so that not less than or more than approximately 100-percent transmission exists at these frequencies.

In order that the low frequencies of the suppressed or lower sideband contribute the proper amount of signal strength to these same low frequencies of the upper sideband for 100-percent transmission, the suppressed lower sideband also has to be attenuated but in a reverse manner. From Fig. 2-4 it is seen that the lower sideband (that portion of the channel to the left of the video carrier) consists only of the video frequencies to 1.25 mc, and most of these low frequencies are at the same (maximum) amplitude. Consequently, the attenuation of the (suppressed) lower sideband will be such that as the low frequencies are increased the attenuation also increases. This is seen from the sloping curve of Fig. 2-6 to the left of the video carrier where the video frequencies of the lower sideband are attenuated in an increasing manner.

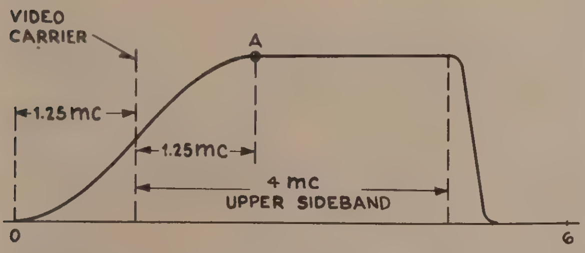


Fig. 2-6.—The over-all bandwidth characteristics of the video channel.

This over-all method of attenuation of the low video frequencies should be such that, if a video frequency, of say, 0.75 mc is attenuated 15 percent in the upper sideband, then that same video frequency of the lower sideband should be attenuated 85 percent in the (suppressed) lower sideband. This means the over-all video frequency of 0.75 mc passed to the video amplifiers will be equivalent to 100-percent transmission. In other words the upper sideband *contributes* 85 percent of the video frequency and the (suppressed) lower sideband contributes 15 percent, making it 100 percent over-all. The same analysis holds for the other video frequencies from zero to 1.25 mc. The video frequencies past 1.25 mc (which are all at the same level) are contained *only* in the upper sideband and, therefore, do not have to be attenuated. In effect we are putting the low and high video frequencies on the same level so that upon reproduction there will be no discrimination between the video frequencies.

The Carrier and Intermediate Frequencies

We have already stated that due to FCC standards the sound carrier for each television channel is located 0.25 mc below the upper frequency limit of the channel. The video carrier should be located 4.5 mc below the sound carrier. This means that the video carrier frequency is 4.75 mc lower in frequency than the upper limit of the television channel or 1.25 mc higher in frequency than the low-frequency limit of the channel. Thus for each one of the 13 television channels the sound and video carrier frequencies have a fixed position in the channel and are 4.5 mc apart. In other words, no matter what service area is covered by the television transmitter the sound and video carrier frequencies for every number 1 channel are the same, those for every number 2 channel are the same, and so on.

Below is a tabulation of each of the 13 television channels with the video and sound carrier frequency used in each channel.

Most manufacturers include the tabulations below in their television service data but they also provide a separate tabulation of the oscillator frequencies for their particular receiver. This oscillator frequency is a fixed but different value for each channel, but for every channel it is used in conjunction with both the video- and sound-modulated signals.

CHANNEL NO.	CHANNEL FREQUENCY (MC)	VIDEO CARRIER FREQUENCY (MC)	SOUND CARRIER FREQUENCY (MC)
1	44—50	45.25	49.75
2	54—60	55.25	59.75
3	60—66	61.25	65.75
4	66—72	67.25	71.75
5	76—82	77.25	81.75
6	82—88	83.25	87.75
7	174—180	175.25	179.75
8	180—186	181.25	185.75
9	186—192	187.25	191.75
10	192—198	193.25	197.75
11	198—204	199.25	203.75
12	204—210	205.25	209.75
13	210—216	211.25	215.75

The front-end section of the television receivers today have a response characteristic the width of a television channel—namely 6 mc. This front end usually consists of an r-f stage, an oscillator stage, and a mixer stage for frequency conversion action. This complete front end functions for both the video and sound signals. Since both the video and sound carrier signals have a fixed amount of frequency separation, one oscillator frequency setting is used to produce both the video i.f. and sound i.f. (The front end is discussed in greater detail in chapter 4.) Once the oscillator frequency is known, it is a simple matter to find out what the center frequency of the video and sound i-f signals will be. The video and sound i.f.'s are not fixed in frequency for each channel, which means that the oscillator frequency of each receiver is determined by the i.f.'s used. However, for every television receiver there is a fixed frequency difference of 4.5 mc between the center i.f. of the video and sound signals. This difference is predetermined by the transmitter which has the video and sound carrier frequencies separated by 4.5 mc as specified by the FCC.

The center frequency of the i.f.'s has been found to vary anywhere from about 25 mc to 27 mc with the accompanying sound i.f.'s falling 4.5 mc below the video.

It appears that there is no definite standardization of the i.f.'s as used in today's television receivers since there are so many different ones in use. The television field is relatively new and perhaps in time to come some standardization may be made on the i.f.'s used. For the sake of the serviceman it is hoped that the *near* future will see some such standardization.

Some examples of typical sound and video i.f.'s are included in the listing below.

SOUND I.F.	VIDEO I.F.
21.25	25.75
21.6	26.1
21.75	26.25
21.9	26.4
22.25	26.75

In all of the above instances, the oscillator frequency tracks above the incoming television signal. When the oscillator frequency beats with the television signal in the mixer tube, a number of signals of different frequencies result. The two frequencies that we are interested in and for which special tuned circuits are employed are the so-called video i.f. and sound i.f. The video i.f. produced is quite wide, and when the output video i.f. of the mixer tube is mentioned, we mean that frequency which is the difference between the video carrier and the receiver oscillator frequency. The same situation holds true for the sound i-f signal.

The reason for this clarification of the video i.f. is that the video i-f transformers used in almost all the television receivers are not tuned to the video i.f. produced by the mixer tube. The video i-f transformers are in most cases stagger tuned with each adjacent i-f transformer, and in many instances all i-f transformers, tuned to a different frequency. When these frequencies are mentioned they represent only the tuned frequency of the video i-f transformers and not the video i-f signal produced by the mixer tube. In chapter 6 the video i-f stages will be discussed in detail along with the theory of stagger tuning.

CHAPTER 3

TELEVISION RECEIVING ANTENNAS

By Seymour D. Uslan

There is no doubt that the television receiving antenna is today considered a vital part of the receiver. There has been many an argument launched as to the necessity of a receiving antenna, but due to the nature of the transmitted television signal, the surrounding terrain, and the location or site of the receiver a receiving antenna is generally considered a must in television.

Because of the relative importance of this component for the *proper* reception of television signals, a separate chapter is devoted to it. In this chapter we will discuss the more common types of receiving antennas and such associated topics as impedance matching, resonance, folded dipoles, reflectors, ghosts and so forth.

The Transmitted Television Signal

The nature of the transmitted television signal is quite different from that of radiobroadcast signals because of the frequencies involved. The television frequencies are short-wave signals, and as such their propagation characteristics differ greatly from the longer wavelength signals in the a-m broadcast bands. These longer wavelength signals follow the curvature of the earth (ground waves) and also use the atmosphere (sky layers) as a means of propagation. When these signals use the sky layers as a means of propagation, they are reflected or refracted from them and thus find their way back to the earth.

The antennas employed in regular a-m broadcasting, therefore, radiate the signal in all directions and do not confine it to any one particular plane. At the higher frequencies (shorter wavelengths) such as those used in television and f.m., the signals are not able successfully to utilize the sky layers or curvature of the earth as a means of travel. Due to the high-frequency nature of these signals, they will pass through the sky layers without being reflected. If the signals try to follow the curvature of the earth, the ground will attenuate them very rapidly; the higher the frequency the greater will be the attenuation.

In order that the television signal reach the receiver with sufficient strength, the signal radiated from the television antenna is primarily confined to the horizontal plane. The necessary service area should be covered with the greatest amount of field intensity in a 360° horizontal plane. A certain amount of vertical radiation is unavoidable, but this radiation is kept to a minimum.

Aside from the direction of radiation of the transmitted signal, there is the problem of *polarization* to consider. When used with reference to radio waves, polarization means the direction of the *electric field* of the signal. The radiated signal is an electromagnetic signal, which means there are two alternating fields involved, the electric field and the magnetic field. These fields are 90° out of phase with each other. When polarization is discussed, reference is made only to the electric field. This means that the signal can be either horizontally or vertically polarized, depending upon the position of the antenna. When the radiating element is horizontal, the electric field alternates in a lateral motion such that the signal is *horizontally polarized;* and when the radiating element is in the vertical plane, the electric field alternates in an up-and-down motion such that the radiated signal is *vertically polarised.* Although it has not been standardized by the FCC, horizontal polarization is used by practically all the television stations in the United States.

There are two good reasons why horizontal polarization is preferred to vertical polarization. In the first place, with vertical polarization much greater losses in the radiated wave are caused by ground and atmospheric attenuation than with horizontal polarization. Secondly, many types of interfering signals are vertically polarized. If the radiated television signal were vertically polarized, the interfering vertically polarized signals would be easily picked up by the receiving antenna.

Television broadcasting stations are generally located approximately in the center of their service areas, so that the horizontal radiation pattern has to be somewhat circular in shape in order to cover the necessary area.

Horizon Range and Line of Sight

Television receiving antennas cannot be as simple as those employed for the longer wavelength signals and still as efficient because of the directive nature of the television signal. Since the propagated television signals have their energy concentrated in one plane, the receiving antenna should be located at a site where it will intercept the signal.

Besides the actual design of the receiving antenna, two other factors play an important part in the proper reception of television signals: the height of the antenna above ground and its distance from the transmitting antenna. In determining these, the *horizon range* and the *line of sight* must be considered. By horizon range is meant that distance covered by the direct wave from the antenna to the curvature of the earth. The line of sight is the maximum distance that can be seen from the top of the transmitting antenna to the top of a receiving antenna before the earth blocks out the view.

Besides the distance between the transmitting and receiving antennas, the height of these antennas also determines the horizon range and line of sight. Fig. 3-1 makes this somewhat clearer. Points *A* and *B* represent two antennas, *A* a transmitting and *B* a receiving antenna. According to the above definitions, line *AX* represents the horizon range from the transmitting antenna and line *AB* represents the line of sight between the two antennas. If the height of the transmitting antenna were raised from *A* to *C*, and if an observation from the top of point *C* were then made, the horizon range would be seen to have increased from distance *AX* to distance *CY*. The line of sight of the transmitting antenna, now at point *C*, is also said to increase. By this is meant that, if antenna *B* is moved so that from its top to the top of antenna *C* a new line of sight is formed, this new line of sight will be greater than *AB*. This is shown in Fig. 3-1, where point *D* represents the new position of antenna *B* and the new line of sight is represented by line *CD*.

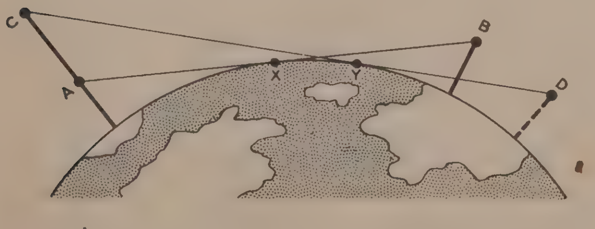


FIG. 3-1.—The line of sight is shown between two antennas at *A* and *B*. If the antenna *A* were raised to position *C*, the line of sight would be extended to *D*.

If a receiving antenna is located in a position where it is beyond the horizon range of the transmitter and does not fall within the line of sight of the transmitting antenna, reception of the signal will, in most instances, not take place. In other words, since the signal is directional and cannot successfully use the atmosphere or ground as a means of propagation, it will be blocked by the earth and not be able to reach the receiving antenna.

The above analysis indicates that in order to increase the possibility of signal pickup, the horizon range and, hence line of sight should be increased. This increase can be achieved if either the transmitting or receiving antenna is raised. In practice the transmitting antenna is made as high as is possible in the service area it is supposed to cover. The receiving antennas should likewise be as high as possible in order to prevent the signal from being blocked by hilly terrain or other intervening objects that effectively cut down the horizon range.

The actual distance of the horizon range and the line of sight can be very easily calculated. The line of sight can be considered as consisting of two horizon ranges—one from the transmitting antenna and the other from the receiving antenna. Thus if we have a method of calculating each horizon range, we can add them together and obtain the line of sight distance. This can be easily seen from Fig. 3-2. The horizon range of the transmitting antenna in this drawing is equal to distance *AB*, and at point *B* it is tangent to the earth. This means that at point *B* it makes a right angle with the radius of the earth drawn from this point. This radius is indicated by line *BC*. Any other straight line drawn from the circumference of the earth to its center is also a radius of the earth. Another such radius is line *CD*. The height of the antenna is represented by distance *AD* and when this line is extended to the center of the earth at point *C* it completes line *AC*. Lines *AB*, *BC*, and *AC* form a right triangle. With the height of the antenna known and the radius of the earth likewise known, it is very easy to calculate the horizon range.

In any right triangle the two sides that form the right angle, as sides *AB* and *BC* in Fig. 3-2, are called the legs of the triangle. The side connecting the two legs, side *AC*, is called the hypotenuse of the triangle. The hypotenuse of any right triangle is always larger than either one of its legs but smaller than the arithmetical sum of the two legs. An old geometrical theorem about right triangles, called the Pythagorean theorem, states that the square of the hypotenuse is equal to the sum of the squares of its two legs. Referring to Fig. 3-2 the theorem is expressed symbolically as follows:

$$(AC)^2 = (AB)^2 + (BC)^2 \qquad \text{Eq. 3-1}$$

Calling the radius of the earth R and the height of the transmitting antenna t and the horizon range distance AB by the letter X, the above relation reduces to the following:

$$(t + R)^2 = X^2 + R^2 \qquad \text{Eq. 3-2}$$

which reduces to the following:

$$t^2 + 2tR + R^2 = X^2 + R^2 \qquad \text{Eq. 3-3}$$

and subtracting the R^2 from either side of the equation we have:

$$t^2 + 2tR = X^2 \qquad \text{Eq. 3-4}$$

The radius of the earth, R, is approximately 4000 miles. Compared with this the height of the antenna, t, even when squared will be very much smaller than the expression $2tR$ and therefore the term t^2 can be neglected in the last equation. Consequently, the final resultant equation is:

$$X^2 = 2tR \qquad \text{Eq. 3-5}$$

or

$$X = \sqrt{2tR} \qquad \text{Eq. 3-6}$$

Since R equals approximately 4000 miles, we can further simplify equation 3-6 as follows:

$$X = \sqrt{2 \times t \times 4000} = \sqrt{8000t} = 88.6\sqrt{t} \qquad \text{Eq. 3-7}$$

Equation 3-7 will give the horizon range X in units of miles when the antenna height t is also given in units of miles. Since the antenna height is usually given in terms of feet, equation 3-7 can be changed to:

$$X = 1.23\sqrt{t} \qquad \text{Eq. 3-8}$$

where X is equal to the horizon range in miles and t the antenna height in feet.

The same type of analysis will hold with the receiving antenna. That is, the horizon distance in miles from the receiving antenna is equal to the product of 1.23 and the square root of the height of the receiving antenna in feet.

As mentioned before, the maximum distance of the line of sight is equal to the sum of the horizon ranges of the transmitting and receiving antennas. Thus, with height of the transmitting antenna designated as t and that of the receiving antenna designated as r the line of sight distance, call it S, is equal to the following:

$$S = 1.23\sqrt{t} + 1.23\sqrt{r} \qquad \text{Eq. 3-9}$$

or we can write equation 3-9 in simpler form as follows:

$$S = 1.23\,(\sqrt{t} + \sqrt{r}) \qquad \text{Eq. 3-10}$$

From equation 3-10 it can be seen that raising the height of either t, the transmitting, or r, the receiving, antenna will increase S, the line of sight.

For example, if the height t of the transmitting antenna is 1000 feet and if the receiving antenna, which falls beyond the horizon range of the transmitting antenna, has a height r of 50 feet, the line of sight S would be calculated as follows:

$$S = 1.23\,(\sqrt{t} + \sqrt{r})$$
$$S = 1.23\,(\sqrt{1000} + \sqrt{50})$$
$$S = 1.23\,(10\sqrt{10} + 5\sqrt{2})$$
$$S = 1.23\,(31.62 + 7.07)$$
$$S = 1.23\,(38.69)$$
$$S = 47.59 \text{ miles}$$

Thus the line of sight between the two antennas is equal to 27.64 miles. If the receiving antenna is raised to 200 feet the line of sight would also increase. Thus:

$$S = 1.23\,(\sqrt{1000} + \sqrt{200})$$
$$S = 1.23\,(10\sqrt{10} + 10\sqrt{2})$$
$$S = 1.23\,(31.62 + 14.14)$$
$$S = 1.23\,(45.76)$$
$$S = 56.28 \text{ miles}$$

Increasing the receiving antenna height from 50 to 200 feet has thus increased the line of sight distance by approximately 8.7 miles.

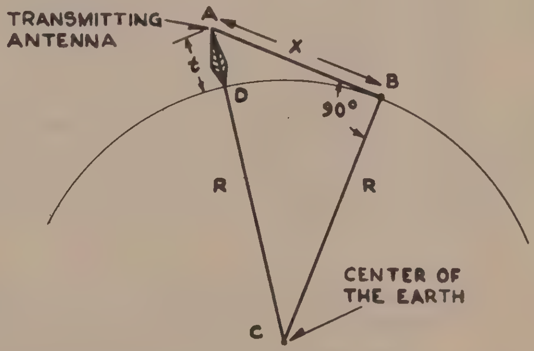


FIG. 3-2.—The horizon range is the distance between points A and B. The horizon range from the transmitting antenna and the horizon range from the receiving antenna can be added together to obtain the line of sight.

Propagation of television signals can cover distances beyond the line-of-sight range when the effective radiated signal from the antenna is strong (that is, either a high-powered transmitter or a transmitting antenna with a high power gain, or both). This is primarily caused by the directed radiated television signal undergoing diffraction and refraction as it travels toward the receiver. This is a result of the nature of the atmosphere, principally the troposphere, that part closest to the earth. Because the signal undergoes diffraction and refraction, it has a tendency to follow the curvature of the earth through the atmosphere and thus cover distances beyond the line of sight range.

The Television Signal at the Receiver

It has often been said that a television receiver is only as good as its antenna. But some authorities have also held that a piece of wire will serve as an antenna, or even that no antenna is needed. These contradictory statements arise from the fact that good television reception has been attained using just straight pieces of wire, and thus, the necessity of a television antenna is disputed. However, the only cases of such reception are of the particular kind where the television receiver happens to be located in a very favorable region with respect to some of the broadcasting antennas. The signal strength is then relatively high, and good reception may be obtained without the use of a television receiving antenna.

If a rigid test is made in such an area with a perfectly matched antenna system, the reception of the television signal *will* be better than without any antenna at all. In practically all installations it can be shown that television reception is definitely improved with the proper type of antenna. At points which are a considerable distance from, but within the service range of the transmitter, the signal pickup of a conventional single wire is usually so small that good reception cannot be obtained.

If you are in doubt whether to use a television antenna in your location and if you do not know too much about the surrounding terrain and where the television broadcasting antennas are located, then it is definitely advisable to use a television receiving antenna.

The nature of the transmitted television signal is a criterion in orienting the television receiving antenna in order to obtain the maximum amount of signal pickup for proper reception. As mentioned previously, the transmitting antennas of most television stations employ horizontal polarization and thus the signal is horizontally directed.

This necessitates the orientation of the television receiving antenna in a horizontal position for maximum signal pickup from the passing television waves. More will be said later about the orientation of the television receiving antenna.

The antenna generally employed for television reception is a single half-wave dipole or some combination of half-wave dipoles, with or without reflector elements. All types of television receiving antennas are designed to perform two main functions, namely, obtaining the necessary signal pickup and supplying an output impedance that can be properly matched to the receiver for maximum energy transfer. The different types of receiving antennas perform these functions in one manner or another, as will be seen later.

When we refer to a signal-to-noise ratio at the input to the television antenna, the noise factor does not include any tube or thermal noise relative to the television receiver circuits. Actually, therefore, two noise components have to be taken into account when the television signal reaches the detector circuits of the receiver. One of the components results from the noises and interfering signals picked up directly by the antenna, and the other is attributed to the inherent noise characteristics of the components within the receiver itself. Since it is difficult to reduce the inherent noise within the receiver, the best thing to do is either to reduce the noise pickup by the antenna system or to increase its signal pickup, so that the signal-to-noise ratio of the television signal just before the detector stages of the television receiver will be high.

To understand why a half-wave dipole is used as the basis of the television antenna, it is necessary to understand something about the voltage and current distribution, antenna resistances, resonance conditions, and the like with respect to such types of antennas.

Voltage and Current Distribution

In choosing the type of antenna for television reception, as well as other types of reception, the voltage and current distribution along the antenna must be known for an understanding of the reason a certain antenna is used. In this respect, let us examine the full-wavelength straight-wire antenna as seen in Fig. 3-3. The distribution is such that at the ends of the wire the current is a minimum and the voltage a maximum. Due to certain characteristic phenomena of open wires used as antennas, such as that appearing in Fig. 3-3, *standing waves* of both voltage and current exist along these antennas. By standing waves is meant the following: A radio-frequency wave traveling to the ends of the antenna will be reflected back to its starting point. The reflected wave will meet the original (so-called incident wave) in such a manner that the individual voltage and current curves of these waves will add algebraically such that voltage and current waves always exist on the antenna. These waves are referred to as *standing waves*. Antennas in general are referred to as radiating elements, or a series of such elements, and for the antenna to be a good radiating element (or receiving element), standing waves should exist on it.

Fig. 3-3 shows the standing wave distribution for a full-wavelength antenna. Distribution of the voltage and current is sinusoidal. Since the voltage and current are represented by sine waves, it is evident that they change polarity at certain points along the full-wavelength wire. The current is a maximum and the voltage zero at one-quarter of a wavelength from either end of the wire. Since the full wavelength of wire represents one complete cycle (360 degrees of electrical length of either voltage or current), the standing waves of voltage and current are said to be 90° out of phase. This 90° phase difference is the same as saying that the reversals of polarity of the voltage and current waves occur one-quarter of a wavelength apart.

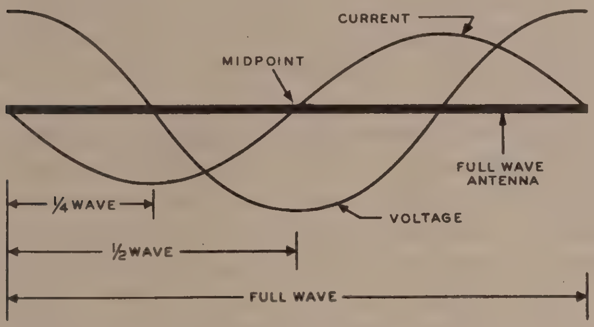


Fig. 3-3.—A full-wavelength straight-wire antenna with standing current and voltage waves. Note that at the midpoint of the antenna the current is a minimum and the voltage a maximum.

The numerous qualities of an antenna apply to both transmitting and receiving antennas, so that those properties discussed under receiving antennas apply equally well to transmitting antennas and vice versa. Consequently, when we talk about the voltage and current distribution of the standing waves on one type of receiving antenna, it applies equally well to the same antenna when used for transmitting purposes.

However, the transmission line used with the antenna should have no standing waves on it, so that the line will not radiate any of the energy but rather transfer it to the receiver. If standing waves do exist on the transmission line, some of the energy picked up by the antenna will be lost from the line as this energy is transferred to the television receiver input and also "ghosts" may appear on the picture tube screen. This will be discussed in detail later on in this chapter. In order to make the transmission line a non-radiating element having no standing waves (that is, no reflection of the signal), the line must be terminated in its characteristic impedance. This means that the impedance of the transmission line must be properly matched to the impedance of the television receiving antenna and also to the input of the television receiver. The standing wave ratios quoted by antenna manufacturers include the loss introduced by the transmission line; the lower the standing wave ratio, therefore the better the antenna and transmission line setup.

The (Half-Wave) Dipole Antenna

Let us now consider a half wavelength of wire or just half that in Fig. 3-3. This half-wave wire is illustrated in Fig. 3-4 and is generally representative of the current and voltage distribution of the half-wave dipole antennas used with many television receivers. A half-wave antenna is often referred to as just a dipole antenna. The terminology (dipole) for these antennas has originated from the fact that voltage distribution along the half-wave antenna is such that at the ends of the antenna the voltages are of opposite polarity (that is, positive and negative charges). This is readily evidenced by the half cycle of voltage in Fig. 3-4. Since the half-wave dipole antenna is the basic type used for television receivers, let us examine some of its characteristics as shown in Fig. 3-4. At any one instant, the current standing wave is of the same polarity at all points, and there is, ideally, zero current at the antenna ends with maximum current, which is referred to as a current loop, at the center of the dipole at one-quarter of a wavelength from either end. The voltage changes polarity, and the change is such that there is, ideally, zero voltage at the center of the dipole, one-quarter wavelength from either end.

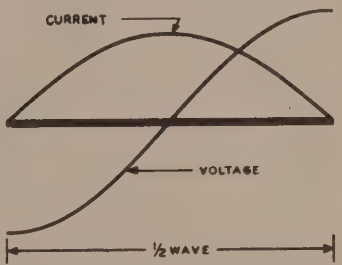
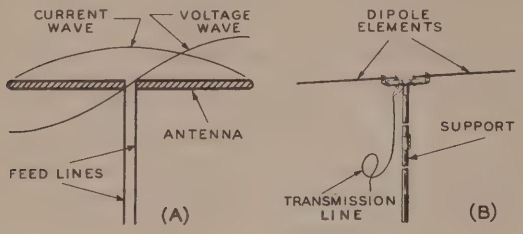


Fig. 3-4.—The current and voltage distribution along a half-wave antenna. Note that at the ends the current is a minimum and the voltage a maximum. Compare with Fig. 3-3.

At the ends of the dipole, the voltage is a maximum, but the voltage at one end is of opposite polarity to the voltage at the other end.

A drawing of a half-wave dipole antenna is illustrated in Fig. 3-5 (A) along with the waves of voltage and current distribution. In Fig. 3-5 (B) is a picture of a typical half-wave dipole antenna as used today. The half-wave dipole antennas illustrated are those that have center lead-ins and are termed *current-fed* antennas, because the feed-in lead is at a current loop or maximum. Half-wave antennas can also be end-fed, but television receiving antennas are generally center-fed dipoles. For proper reception, the half-wave dipole antenna should have a length *approximately* equal to one-half of a wavelength at the center frequency of the frequency spectrum it is to receive. Consequently, the length of each section should be about one-fourth of the wavelength of the middle frequency which the antenna is designed to receive. This will be elaborated upon later in this chapter.



(B) Courtesy Technical Appliance Corp.

FIG. 3-5.—Current and voltage distribution in a half-wave antenna (A). In (B) is shown a typical dipole antenna, its transmission line, and a method of support.

Antenna Resistances

All the power input to a transmitting antenna is dissipated in one form or another. The so-called *resistances* of an antenna determine how this power is dissipated. Any antenna, whether it is used for receiving or transmitting, effectively contains two types of antenna resistance. One type is the usual ohmic resistance of the metal parts of the antenna, often called the *real resistance;* the other type is often called the imaginary or *radiation resistance*. In other words, we are concerned with two types of power, the power dissipated due to the actual ohmic resistance of the antenna and the power radiated from the antenna. The former power is readily understandable and, from this, the ohmic resistance is conceivable. The resistance dissipating the latter portion of the power is not, in reality, a physical resistance in the sense that the other one is and, hence, often is known as the imaginary or radiation resistance of an antenna. Consequently, when there is talk of power dissipation of an antenna, the I^2R total power loss should be understood to encompass both types of resistances. Resistance R is the series combination of the ohmic resistance of the antenna and the radiation resistance of the antenna.

In most types of antennas, the ohmic resistance is much smaller than the radiation resistance, so that practically all the power is dissipated through the radiation resistance. For a half-wave dipole antenna in free space (that is, no intervening objects including ground effects, buildings, mountains, etc.) the radiation resistance is found to be equal to approximately 73 ohms. The actual value of radiation resistance of the simple dipoles, as used, varies somewhat away from 73 ohms, depending on the exact length of the dipole and presence of other physical factors.

Resonance and Impedance

The half-wave antenna behaves very much like a tuned circuit. The antenna has inductive, capacitive, and resistive components, but they are all distributed throughout the antenna in contrast to the lumped L, C, and R components in the regular tank circuit. The simple half-wave dipole antenna that is center tapped behaves like a series resonant circuit. This is shown in Fig. 3-6 (B) where L, C and R represent the inductive, capacitive, and resistive components respectively. The resistive component is primarily the radiation resistance of the antenna since the real or ohmic resistance is relatively very small, even at the increased value it has at high frequencies.

If the dipole is approximately a half wavelength long, or about one-quarter wavelength on either side of the center tap, the antenna will act as a series resonant circuit to the frequency for which it is a half-wave antenna. In a series resonant circuit the impedance is a minimum and purely resistive, the current a maximum, and the voltage a minimum. This is also seen from Fig. 3-6 where, at the center point of the half-wave antenna, the current is a maximum and the voltage a minimum. At other points along the antenna the impedance is not purely resistive because it encompasses some reactance. This is due to the changing relationship of the voltage and current standing waves and, therefore, the impedance is greater away from the center point, being a maximum at either end of the half-wave antenna due to the minimum value of current and maximum value of voltage.

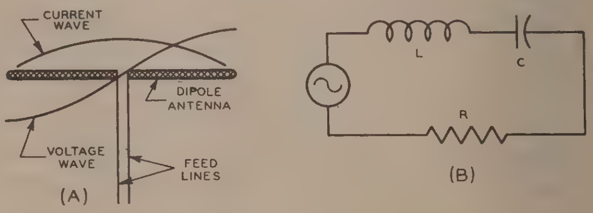


FIG. 3-6.—A half-wave dipole antenna (A) and its equivalent series resonant circuit (B), where the resistive component R is primarily the radiation resistance of the antenna.

Transmission Lines

The transmission line connecting the antenna to the receiver is as important a factor in obtaining the correct input signal to the receiver as the antenna itself. The transmission line is primarily intended to provide a ready path for the television signal from the antenna to the receiver, with a minimum amount of signal loss. To adequately provide this ready path, the transmission line must *match* the impedance of the antenna to the input impedance of the television receiver. For this so-called impedance match, which is discussed in detail in the following section, the *characteristic impedance* of the transmission line has to be known.

No matter what type of transmission line system is used, each one is distinguished by its own characteristic impedance which is expressed in ohms. There are two main types of transmission lines used with today's television receiver—the open-wire line, sometimes also called twin-lead, twin-pair, or ribbon type, and the coaxial line. The characteristic impedance, sometimes called surge impedance, iterative impedance, or just plain impedance is determined by the physical make-up of the line and can be calculated by knowing the inductance and capacitance per unit length of line (assuming a lossless line in the latter case.)

The impedance value of a transmission line is not a direct function of the length of the line, although this is often thought to be the case. In other words, if the known characteristic impedance of a transmission line is equal to, say, 100 ohms, this value of impedance does not refer to any specific length line but to the *type* of line itself. That is, 500 feet of the line will not have twice as much (characteristic) impedance as 250 feet of the same line but will have the *same* characteristic impedance.

If a transmission line had a voltage applied across its input terminals, the ratio of this voltage to the current at the input end would be the input impedance. If the line were uniform and of infinite length, the input impedance would become the *characteristic impedance* of the line. Thus if a line infinitely long had a 50-foot section cut off, the characteristic impedance of the line would not change, because subtracting a finite length of line from one that is infinitely long still leaves the line infinitely long. The piece cut out is a part of the infinitely long line and as such the characteristic impedance is a means of identifying this part. The input impedance of this 50-foot line may be different from the characteristic impedance of the infinitely long section. However the 50-foot piece is still said to have the same characteristic impedance as the infinite line.

If the finite line were terminated in its characteristic impedance, we would not be able to tell by simply measuring the voltage and current at the input end whether the line was of a finite length or infinitely long. In other words, when a finite line is terminated in its characteristic impedance, the input impedance of that line is equal to the characteristic impedance. When this fact is known, maximum energy transfer from the antenna to the receiver via the transmission line is easily attained. This will be seen in the next section on impedance matching.

If the leakage conductance and series resistance per unit length of line is negligible, as is usually the case, the characteristic impedance of a line is purely resistive and can be represented by a relationship between the inductance L and capacitance C per unit length of line. Denoting the characteristic impedance by the common expression Z_o, it is represented mathematically as follows:

$$Z_o = \sqrt{\frac{L}{C}} \qquad \text{Eq. 3-11}$$

This expression holds for both twin-lead transmission lines and the coaxial type. The characteristic impedance can also be given in terms of the physical characteristics of each type of line. For the twin-lead transmission line, as seen in Fig. 3-7, the characteristic impedance Z_o is given by the following:

$$Z_o = \frac{276}{\sqrt{\epsilon}} \log_{10} \frac{2D}{d} \qquad \text{Eq. 3-12}$$

where D is the distance between centers of the wires
d is the diameter of the wire (whether one wire or a number of small wires grouped together)
ϵ represents the dielectric constant of the material surrounding the wires. For air $\epsilon = 1$ and for polyethylene $\epsilon = 2.25$. (This value may vary slightly.)

(Equation 3-12 is obtained from equation 3-11 by substituting the expressions for L and C for the parallel transmission line of Fig. 3-7.)

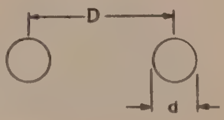


FIG. 3-7.—Cross-sectional view of a twin-lead transmission line. D is the distance between centers of the twin leads and d is the diameter of each lead.

In using such a formula for computation, the units of D and d should be the same, whether they are measured in inches or centimeters. After computing $2D/d$ the logarithm of the answer ($\log_{10}$) can be found in any book containing tables of logarithms. Remember, however, that $\log_{10}$ means "logarithm to the base 10" and refers to a table of "common logarithms" and not natural logarithms, which are computed to a different base.

For example, let us compute the characteristic impedance of a parallel-wire transmission line that is covered by polyethylene plastic insulator ($\epsilon = 2.25$) and measures 22/64 of an inch between centers with the wire having a diameter of 3/64 of an inch. Using equation 3-12 for the characteristic impedance Z_o we have:

$$Z = \frac{276}{\sqrt{\epsilon}} \log_{10} \frac{2D}{d}$$

and since $\epsilon = 2.25$, $D = 22/64''$, and $d = 3/64''$

$$\begin{aligned} Z_o &= \frac{276}{\sqrt{2.25}} \log_{10} \frac{2(22/64)}{3/64} \\ &= \frac{276}{1.5} \log_{10} \frac{44}{3} \\ &= 184 \log_{10} 14.66 \\ &= 184\ (1.662) \\ &= 306 \text{ ohms} \end{aligned}$$

Such a transmission line would be listed as having a 300-ohm characteristic impedance.

Referring to Fig. 3-8, the characteristic impedance for the coaxial transmission line is as follows:

$$Z_o = \frac{138}{\sqrt{\epsilon}} \log_{10} \frac{D}{d} \qquad \text{Eq. 3-13}$$

where D is the inside diameter of the outer conductor (or the outside diameter of the dielectric)

d is the diameter of the inner conductor (whether one wire or a number of small wires grouped together)

ϵ represents the dielectric constant of the dielectric material. For air $\epsilon = 1$, for polyethylene $\epsilon = 2.25$. (This latter value may vary slightly).

(Equation 3-13 is obtained from equation 3-11 by substituting the expressions for L and C for the coaxial transmission line of Fig. 3-8.) In equation 3-13, as well as equation 3-12, both D and d have to be in the *same* units.

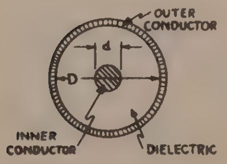


FIG. 3-8.—Cross-sectional view of a coaxial transmission line. D is the inner diameter of the outer conductor and d is the outer diameter of the inner conductor.

The method of using equation 3-13 to compute the characteristic impedance of a coaxial transmission line from its physical dimensions is similar to the example just discussed for the parallel wire. For example let us compute Z_o for a coaxial line that uses polyethylene as a dielectric ($\epsilon = 2.25$), where the diameter D of the dielectric and the diameter d of the inner conductor are 19/64 and 3/64 of an inch respectively. Thus

$$\begin{aligned} Z_o &= \frac{138}{\sqrt{\epsilon}} \log_{10} \frac{D}{d} \\ Z_o &= \frac{138}{\sqrt{2.25}} \log_{10} \frac{19/64}{3/64} \\ Z_o &= \frac{138}{1.5} \log_{10} \frac{19}{3} \\ &= 92 \log_{10} 6.33 \\ &= 92\ (.8014) \\ &= 73.7 \text{ ohms} \end{aligned}$$

Such a coaxial line is often said to have a 75-ohm impedance.

The parallel wire and coaxial type of transmission lines are available in various characteristic impedances. Most antennas used in television reception have either about a 75-ohm impedance as the simple half-wave dipole antenna or about 300 ohms as the folded dipole. Consequently, the transmission lines used should have either a 75-ohm or a 300-ohm characteristic impedance. When the coaxial line is used, it is primarily employed for impedance matching between a low-input impedance receiver and a simple dipole antenna. For high-input impedance (about 300 ohms) television receivers, the 300-ohm twin-lead transmission line is usually employed. The above methods are the most commonly used, although variations do exist.

Impedance Matching

So that the maximum amount of energy pickup will be transferred from the half-wave antenna to the receiver input, the antenna has to be properly "impedance-matched" to the receiver input. A fair idea of impedance matching can be had from Fig. 3-9. Looking in the direction of the r-f amplifier, through the primary of the input transformer, we see an impedance equal to Z_1; and looking in the direction of the transmission line and antenna, we see an impedance of Z_2; for maximum energy transfer Z_1 should equal Z_2. Under this condition the maximum amount of energy possible, not *all* the energy, will be transferred from the antenna to the grid of the first tube. For the best match to occur, the transmission line should first be impedance matched to the antenna. Then these units through the medium of the receiver input transformer should be matched to the input impedance of the r-f amplifier. If there is any mismatch, there will be a loss of energy, and the maximum possible amount of energy will not be transferred.

As has been pointed out, the input impedance at the center of the half-wave dipole ideally is a pure resistance and equal to 73 ohms. In practice, this value of input impedance of the dipole can vary anywhere from 50 to 100 ohms as a result of such varying physical factors as the construction of the antenna, the obstacles near the antenna, and its height. A very simple impedance match to such an antenna is made by using a twisted-pair transmission line (usually, ordinary rubber-insulated wire) for the feeder section, as shown in Fig. 3-10. The so-called characteristic impedance of such a transmission line is somewhere around 75 ohms. If different types of wire are chosen for the twisted line, and if the physical hookups to the antenna can be varied somewhat, the impedance of this transmission line can be varied on either side of 75 ohms for the desired impedance match. Beside giving a good impedance match, this type of line minimizes pickup by the lead-in due to the twisting effect and closeness of spacing between each individual wire, and thus noise pickup by the transmission line is reduced.

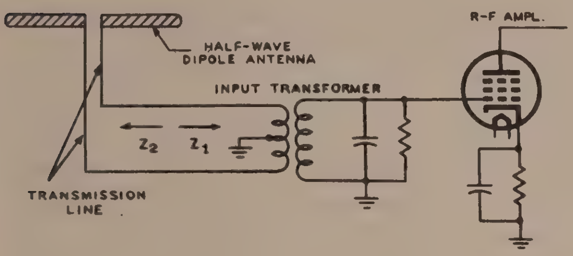


FIG. 3-9.—In order to effect a maximum transfer of energy from the antenna to the control grid of the r-f amplifier tube, the impedance Z_2 of the transmission line and antenna should equal the impedance Z_1 seen looking into the primary of the input transformer.

Where the antenna and line are properly matched, the input transformer of the receiver should have an impedance approximately equal to that of the line or antenna. For example, if the antenna resistance and characteristic impedance of the line are each about 100 ohms, then looking *into* the primary of the input transformer, toward the r-f amplifier, the impedance seen should also be 100 ohms. Under these circumstances, the maximum amount of energy transfer will be made from antenna to receiver. In other words, for maximum energy transfer in any kind of system, the impedance of the load should equal the impedance of the source. If the impedances are not purely resistive (that is, contain reactances) then the reactances in the load and those in the source must be of opposite sign. In television receiving antenna systems, the impedances involved are considered pure resistances for matching purposes.

In reality, almost any good type of balanced transmission line, besides the simple twisted pair, can be used to match the antenna as long as the characteristic impedance of the line is close to the radiation resistance of the antenna. Many receivers use television antennas supplied by outside manufacturers; some receivers are not directly supplied with an antenna, which must be bought separately. In the latter case knowledge of the input impedance of the receiver's input transformer is necessary to secure the proper impedance match. Many television receivers today have an input impedance equal to about 300 ohms, so that the transmission line and antenna have to be properly matched to this 300-ohm impedance for proper energy transfer. Transmission lines with a characteristic impedance of 300 ohms would normally be used to match the line to the transformer, but the 300-ohm line would be mismatched to, for example, a 75-ohm simple half-wave dipole.

Some mismatch can be tolerated under certain conditions. First of all, let it be understood that for maximum efficiency the transmission line should match the impedance of the antenna to the input impedance of the receiver. However, a certain amount of mismatch may be allowable provided that satisfactory performance of the television receiver is obtained. Most television receivers are sensitive enough to suffer a certain amount of energy loss in the received signal and still operate satisfactorily. This means that if there is a small amount of impedance mismatch between the antenna and receiver there will be a certain amount of energy lost by the setting up of standing waves in the transmission line, and if this loss is small, it can be tolerated. Even though mismatching by as high as a 4 to 1 ratio has resulted in so-called satisfactory receiver performance, this type of mismatch is definitely undesirable and should be avoided wherever possible. Besides energy loss such

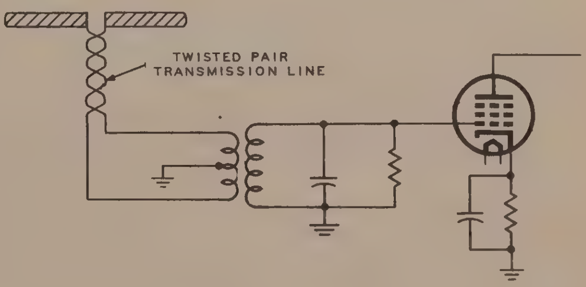


FIG. 3-10.—In order to match the impedance of a half-wave dipole antenna with that of the transmission line, the latter can be formed of twisted pair, which has a characteristic impedance of approximately 75 ohms.

a mismatch can also cause "ghosts" on the picture tube. This will be discussed in detail later on. Since most antenna and transmission line manufacturers list the impedances of their products, it should be quite easy to obtain a match between the antenna and receiver. The difficulty in the complete matching process is in knowing the input impedance of the television receiver. A great many of these receivers have an input impedance equal either to approximately 300 ohms or about 75 ohms, but the impedance of some differs from these. It is readily seen that it is difficult to effect a complete impedance match when it is necessary to guess the input impedance of the receiver.

If the mismatch is too great, there may not be enough signal input to the television receiver to overcome the inherent noise of the receiver. The signal-to-noise ratio will be low, and the reproduced picture signal, as well as the audio signal, will be noisy and incoherent.

It has been stated that for the best reception the line should be matched to the antenna. However, in certain instances simple dipoles of 75-ohm input impedance have been used with 300-ohm transmission lines, yet the signal input to the television receiver is said to be enough for proper reception. With this mismatch of 4 to 1 the set still operates satisfactorily. Many television antenna manufacturing companies stipulate that, if a 300-ohm transmission line is used with their simple half-wave dipole antennas, which have an input impedance around 75 ohms, proper reception will still be attained. Knowledge of the construction, the *Q*, and the impedance of the dipole is necessary to understand completely how such a mismatch can be tolerated.

The input of a simple dipole is, as mentioned, about 75 ohms and resistive in nature. This impedance varies from the center to either end, with the impedance at the ends being a maximum and somewhere in the vicinity of 2500 ohms. Since this is so, if the antenna can be tapped at some point away from the center, where the impedance is about 300 ohms, we have a good place to match a 300-ohm line. This procedure is somewhat impractical, but the same effect is obtained in another way. The 75-ohm impedance that appears at the center of the dipole antenna is measured at the frequency for which the antenna is cut. On either side of this center frequency the impedance of the antenna at the center leads will increase—the amount of increase being determined by the frequency change. Thus an impedance mismatch of 4 to 1 for a 300-ohm line attached to a simple half-wave dipole antenna occurs only at the center frequency for which the antenna is cut; on either side of this frequency the impedance increases and, hence, the impedance mismatch decreases. Therefore, for these off-resonance frequencies the dipole antenna does *not* create a 4-to-1 mismatch with a 300-ohm transmission line. However, this 4-to-1 mismatch still exists at the center frequency to which the antenna is cut, but if the antenna is cut to a frequency which has a relatively strong field in the vicinity of the antenna, the input signal may be strong enough to overcome the loss the mismatch entails and, hence, reception at this frequency may be satisfactory. It should be remembered that the increased impedance on either side of the center of the antenna is not purely resistive but encompasses some reactance.

However, variations of the simple half-wave dipole antenna and variations in transmission line hookups exist, so that impedance matching can be closely attained. There are folded dipoles, dipoles with reflectors, specially constructed simple half-wave dipoles that change the effective radiation resistance, and other unique types of dipole arrangements. Analysis of some of these types of television receiving antennas is made later.

For any combination of transmission line and antenna, the smaller the length of the line, the less will be the line losses, no matter how well it is impedance matched. If the line *must* be long, due to the location of the antenna, low-loss transmission lines should be used to prevent excessive reduction in the signal reaching the receiver.

In conclusion, it may be said that, although correct impedance matching will give the best over-all performance, the primary purpose in most television antenna installations is to obtain the necessary voltage input to operate the receiver satisfactorily.

Q of Antenna

The inductance and resistance of a straight piece of round wire will decrease, as the diameter of the wire increases. In other words, both *L* and *R* are inverse functions of the diameter of the wire. However, at operating frequencies within the television band the inductance decreases at a *faster* rate than the resistance for changes in diameter.

Since the *Q* of a circuit or a wire may be defined as the ratio of inductive reactance to resistance, the *Q* of the dipole is seen to *decrease* as its diameter increases. This is due to the fact that the inductance and, hence the inductive reactance, decreases at a much faster rate than the resistance as a result of the diameter increase. This is the same as saying that the resistance effectively increases with respect to the inductive reactance. Consequently, it can be said that the Q of a dipole antenna is an inverse function of its diameter. The inductive reactance, and hence the Q of a wire, increases with increase in wire *length,* but for the foregoing analysis of the Q of an antenna the length of the dipole, as well as the frequency of operation, is considered constant. The lower the number of wire used for the dipole, the greater will be the diameter and hence the lower the Q.

Since the low-frequency (channels one through six) and the high-frequency (channels seven through thirteen) groups of the television band are quite wide, the antennas used for either or both groups should have as broad a response as possible. This broad response is necessary in order to receive all the signals within each television channel with little or no discrimination. By lowering the Q of the antenna it is possible to secure a fairly broad-band characteristic. However, since a reduction in Q "flattens" out the response characteristic, there is also a reduction in signal pickup. Consequently, a compromise has to be effected between the tolerable loss in signal pickup and the reduction in Q.

The Folded Dipole

The folded dipole antenna has a great advantage over the simple dipole antenna in that it exhibits a much higher impedance, thus allowing for a better impedance match to receivers having 300-ohm inputs. A folded dipole antenna is the simple half-wave antenna illustrated in Fig. 3-5 (A) with another half-wave antenna section joined to it at the ends. A drawing of a typical folded dipole is illustrated in Fig. 3-11. From this drawing the folded dipole is seen to be about a full wavelength of wire (or appropriate tubing or rod) bent so that it takes on approximately the shape shown in Fig. 3-11. In Fig. 3-12 are shown three typical folded-dipole antennas as used today. Those shown in parts (A)and (B) are typical outdoor folded dipoles, while that shown in part (C), although suitable for outdoor use (with proper mechanical support), is especially useful as an indoor antenna. This latter antenna can be placed under a rug, in a closet, or some other convenient place in the home. It is made completely out of 300-ohm twin-lead transmission line; its construction is discussed in a following section on indoor antennas.

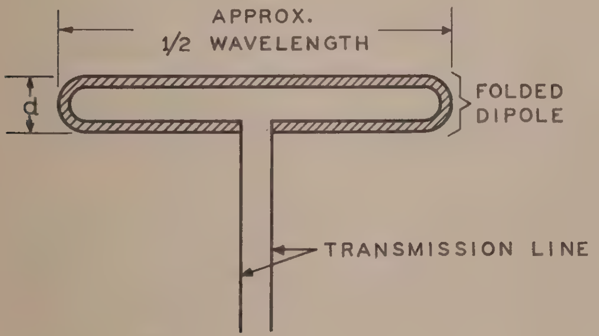
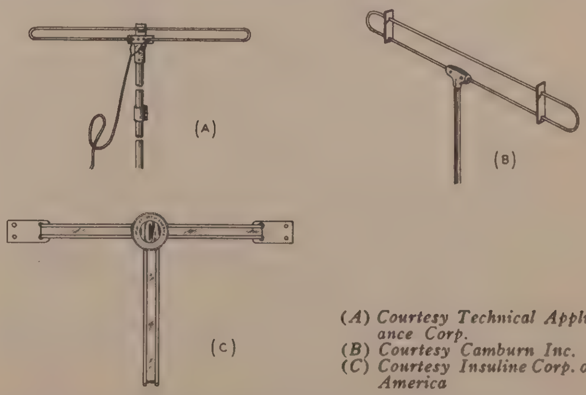


FIG. 3-11.—A folded half-wave dipole antenna, where *d,* the distance between the two half-wave sections, is small in comparison with the wavelength.

Let us now study the folded dipole in general and see how this type of antenna increases the input impedance compared with a simple half-wave antenna. The folded dipole is similar to an autotransformer where the primary of the transformer is analogous to that part of the folded dipole which has the transmission line attached to it and the secondary of the transformer is analogous to the other half-wave section of the folded dipole. Accordingly it is readily seen that a mutual impedance exists between both half-wave sections of the folded dipole in the same way that mutual inductance exists between the windings of a transformer.

Each individual half-wave section of the folded dipole is considered as a resonant half-wave antenna and fundamentally resistive at the resonant frequency. This is the same as having both the primary and secondary of the transformer tuned to the same resonant frequency in which the impedance of each will be resistive in nature. This means that looking into the tapped portion of the folded dipole the complete impedance seen is the sum of the tapped half-wave section plus the *reflected impedance* from the other half-wave section. Reflected impedance means that because of mutual impedance between both half-wave sections of the folded dipole there is an impedance reflected into the primary which changes the over-all input impedance of the antenna. This reflected impedance is mathematically equal to the square of the



(*A*) *Courtesy Technical Appliance Corp.*
(*B*) *Courtesy Camburn Inc.*
(*C*) *Courtesy Insuline Corp. of America*

FIG. 3-12.—Three types of folded dipole antennas. Those illustrated in (A) and (B) are for outdoor use and that in (C) is for indoor or outdoor use.

mutual impedance (which is purely reactive in nature) divided by the impedance of the untapped half-wave section. When multiplying the mutual reactance by itself, the result is a purely resistive component, and since the impedance of the untapped half-wave section of the folded dipole is resistive (at resonance), the reflected impedance is, also, effectively resistive. Consequently the total input impedance of the complete antenna system, as seen from the feed-in half-wave section, under the above conditions of resonance, is resistive.

Since the distance *d* separating the two half-wave sections of the folded dipole as seen in Fig. 3-11 is much smaller than the half-wave length of the antenna, the mutual impedance is considered to approach the maximum possible value. In other words, using the analogy of the transformer again, the coefficient of coupling of the half-wave sections of the folded dipole is said to be approximately unity. Since each individual half-wave section has its own so-called self-impedance, similar to the self-inductance of the windings of a transformer, the total *self-impedance* of both sections (since they are connected) is equal to the sum of their individual self-impedances. The total self-impedance of the folded dipole is not the *complete* impedance, because the folded dipole antenna impedance as a whole takes into account the mutual impedance as well as the total self-impedance. The wire, rod, or tubing used for both half-wave sections is usually made of the same material, so that the self-impedance of both half-wave sections are about the same, neglecting the small difference caused by the slight spacing of the feed-in section of the half-wave element that has the transmission line attached.

For an autotransformer wired in *series aiding,* the total inductance of the unit is equal to the sum of the individual self-inductances *plus twice* the value of the mutual inductance. If the individual self-inductances are equal and if the coefficient of coupling is unity, the value of the mutual inductance will be the same as either self-inductance. Under these circumstances, the total inductance of the autotransformer is equal to four times the self-inductance of one part of the winding.

The same is true, for the most part, of the folded dipole. The individual self-impedances of the half-wave sections are about equal, and the coefficient of coupling between these sections is considered to be unity, so that the mutual impedance is the same as either individual self-impedance. The total input impedance for the folded dipole is, therefore, the sum of the individual self-impedance plus twice the mutual impedance. Consequently, the total input impedance of the folded dipole is equal to four times the value of either self-impedance. For the half-wave folded dipole these impedances are resistive, as previously mentioned, and, thus, the total input resistance of the folded dipole is equal to four times the input resistance of a single half-wave section of the folded dipole. Since a simple half-wave dipole antenna has approximately the same input resistance as a single half-wave section of the folded dipole, the input resistance of a folded dipole is about four times as great as that for a simple half-wave dipole.

Under ideal conditions the input resistance of a half-wave dipole is equal to about 75 ohms. So that under similar conditions the input resistance of the folded dipole is equal to 4 × 75 or 300 ohms. Thus for television receivers having a 300-ohm input impedance, a 300-ohm transmission line can be used to match the folded dipole to the receiver for maximum energy transfer.

Length of the Half-Wave Antenna

Throughout this chapter we have constantly used such nomenclature as the length of the dipole being a half-wavelength long. This was all preliminary to solving the problems of the actual physical length of the antenna and choosing a particular length. If only one frequency is going to be picked up (that is, for a fixed frequency receiver), the antenna length is easy to calculate and is based on that signal frequency. However, we are concerned with television broadcast receivers, and the television band today is essentially from 44 to 88 mc and from 174 to 216 mc, so that the antenna length for either group must be chosen so that it will be responsive to all frequencies within the group.

Most antennas that are in use today are designed primarily for use on the low-frequency television band. This means that the antenna should be broadband to all frequencies in this band, namely 44 to 88 mc. However these antennas are not usually satisfactory when used on the high-frequency (174 to 216 mc) part of the television band. For the high-frequency channels, a different antenna generally has to be used.

Since often some signals are stronger than others, the antenna length should favor the weaker stations. For most practical purposes, however, the length of the antenna is designed for the center frequency of the band of frequencies it is to receive. Accordingly, the dipole antenna for the low-frequency channels is usually cut to 66 mc, the midfrequency of 44 to 88 mc, and for the high-frequency channels it is usually cut to 195 mc. This is based on the assumption that the different television signals in each band are essentially of the same signal strength. However this is not actually so. But in areas where the signal strengths are relatively strong, this analysis can be applied. In those areas where one signal (in either the high- or low-frequency television band) is the weakest of the group, the antenna is often cut to this frequency for maximum possible reception on the channel in question. We will now analyze the method of computing antenna length.

It is known that a 10-meter wavelength means a frequency of 30 mc, but the simple formula telling how this is brought about is often forgotten. The wavelength of a specific frequency is found by dividing this frequency into the *velocity of radio waves.*[1] The velocity of radio waves is equal to 300,000,000 meters per second, and thus with the frequency, f, in cycles per second, the wavelength, in meters, is given by the following:

$$Wavelength = \frac{300{,}000{,}000}{f} \text{ in meters} \quad \text{Eq. 3-14}$$

This is for one full wavelength. If we change the units of this formula and divide the right hand side by 2, we will find that

$$L = \frac{492}{f\ (\text{mc})} \text{ in feet} \quad \text{Eq. 3-15}$$

or

$$L = \frac{5904}{f\ (\text{mc})} \text{ in inches} \quad \text{Eq. 3-16}$$

where L is equal to the length of a *half wavelength* in free space and f is the frequency in megacycles per second.

In half-wave antennas, a so-called "end effect" is attributable to the material supporting the antenna and other physical factors, all of which make the electrical length of a half-wavelength antenna effectively longer than the physical length. To make sure that the electrical length of the antenna used is *effectively* one-half wavelength long, the actual physical length is made less than that for a half wave in free space which has no end effects. The end effect is more pronounced at the higher frequencies, which means that the effective electrical length increases with increase in frequency. From about 5 to 30 mc, the physical length of the half-wave antenna should be reduced by about five percent to make it operate effectively as a half-wave antenna. Since the 44- to 88-mc low-frequency television band has a center frequency of 66 mc, the effective electrical length will increase further, which means the physical length should be reduced by more than five percent. For most practical purposes at the frequencies within the low-frequency television band, the physical length of the half-wave antenna should be reduced by about 6.5 percent. This means that the preceding formulas, equations 3-15 and 3-16, have to be multiplied by 93.5 percent (the difference between 100 and 6.5 percent) to give the correct effective half wavelength.

Thus,

$$L = \frac{492 \times .935}{f\ (\text{mc})} = \frac{460}{f\ (\text{mc})} \text{ in feet} \quad \text{Eq. 3-17}$$

or

$$L = \frac{5904 \times .935}{f\ (\text{mc})} = \frac{5520}{f\ (\text{mc})} \text{ in inches} \quad \text{Eq. 3-18}$$

where L is equal to the effective length of a half-wave antenna.

For most center lead-in half-wave antennas, each half of the dipole is approximately equal to half the values of L found in the foregoing formulas.

With respect to the high-frequency television band (174 to 216 mc), the end effect is even more pronounced. For most practical purposes at these frequencies, the physical length of the half-wave antenna should be reduced by about 8.5 percent. Consequently, equations 3-15 and 3-16 have to be multiplied by 91.5 percent to give the correct effective half wavelength for the high-frequency television band. Thus:

$$L = \frac{492 \times .915}{f\ (\text{mc})} = \frac{450}{f\ (\text{mc})} \text{ in feet} \quad \text{Eq. 3-19}$$

$$L = \frac{5904 \times .915}{f\ (\text{mc})} = \frac{5400}{f\ (\text{mc})} \text{ in inches} \quad \text{Eq. 3-20}$$

Let us now consider a practical example for each television band. Thus for the center frequency of 66 mc of the low-frequency band, the length of a half-wave antenna would be (using equations 3-17 and 3-18) as follows:

$$L = \frac{460}{66} = 6.97 \text{ feet}$$

or approximately 7 feet, and

$$L = \frac{5520}{66} = 83.6 \text{ inches}$$

or approximately 84 inches.

For the center frequency of 195 mc of the high-frequency band, the length of a half-wave antenna (using equations 3-19 and 3-20) would be as follows:

$$L = \frac{450}{195} = 2.31 \text{ feet}$$

or approximately 2.3 feet,

$$L = \frac{5400}{195} = 27.7 \text{ inches}$$

or approximately 28 inches.

---

[1] Many sources in establishing the wavelength of radio waves state that the frequency of the wave is divided into the velocity of *light.* However, since both light waves and radio waves are electromagnetic waves, they basically travel at the same velocity, namely 300,000,000 meters per second. Hence, in the above problem we have omitted the reference to the velocity of light.

If it is desired to compute the antenna length for frequencies other than those at the center of each channel group, equations 3-17 through 3-20 can still be used. All that has to be done is to insert the proper frequency into the denominator of the equations. Channels 1 through 6 use equations 3-17 and 3-18 and channels 7 through 13 use equations 3-19 and 3-20. The frequency that is used for the channel desired is the center frequency of that channel. For channel 1, frequency range 44 to 50 mc, the frequency to use would be 47 mc in equations 3-17 or 3-18. For channel 7, frequency range of 174 to 180 mc, the frequency to use would be 177 mc in equations 3-19 or 3-20.

Since each section of a dipole is effectively a quarter wavelength long, the physical length of each such section would be half that computed in the above examples. However, due to the gap in the lead-in dipole for the transmission line, the actual physical lengths of the individual sections of the dipole are somewhat less than calculated.

All the relations to the length of the half-wave antenna in this section apply equally as well to folded dipoles as to simple dipoles. It has been found in many instances that for best results the length of the folded dipole should be made slightly shorter than its calculated value.

Indoor Antennas

In places where an outdoor antenna cannot be employed with a television receiver, an indoor antenna can be used. There are a few different types of indoor antennas on the market today. One of the most common types used is made of a 300-ohm twin-lead transmission line as indicated in Fig. 3-12 (C). This type can be very easily constructed for any television band or channel desired. A drawing of the construction is illustrated in Fig. 3-13. The letter *L* on the drawing refers to the length of the twin-lead transmission line for the frequency to which it is to be cut. This length *L* can be computed through the use of equation 3-15 or 3-16 and a multiplying factor of 0.825 for both television bands.

For instance, if the indoor antenna is to be cut for the center frequency of the low-frequency television band, then *L* would be equal to about 74 in. Thus a length of 300-ohm twin-lead line equal to a little more than 74 in. is cut; the extra length is used to twist the ends together. The general construction of this indoor antenna is illustrated in Fig. 3-13 and is as follows.

After computing *L,* take a piece of 300-ohm transmission line slightly larger than the computed length and strip away some of the plastic insulator from the ends. Next, twist together and solder the two bare leads on each end, as shown in Fig. 3-13. Then center cut one lead of the twin line and strip away the insulation from each cut end to half the distance between the parallel wires of the transmission line. Finally bend these two bare pieces of wire at right angles at the beginning of the insulator. This part of the antenna represents a folded dipole. Next, take another piece of the 300-ohm twin-lead line and strip off some of the plastic insulator at one end, so that the twin leads are bare. This latter twin lead represents the transmission line, and it is usually any length that will make its wiring to the receiver input most convenient, but the shorter the better. Solder these two bare leads to those of the center part of the folded dipole as shown in Fig. 3-13. The impedance of this folded dipole antenna is about 300 ohms, similar to an outdoor folded dipole.

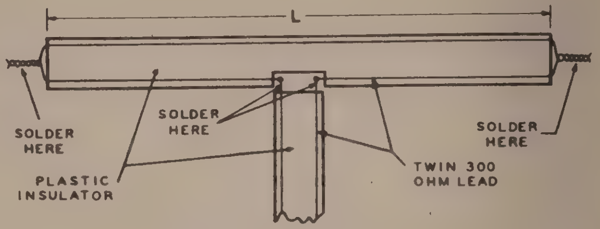


FIG. 3-13.—Construction details of a folded dipole antenna made from a 300-ohm twin-lead transmission line.

Dipole with a Reflector

In many localities the television signal pickup required for proper reception is greater than that obtainable with a half-wave dipole or folded dipole antenna alone, so that something has to be done to increase the signal pickup. This is especially necessary when the receiver is located at a great distance away from the transmitting antennas. Since the signal surrounds the antenna, it is easily realized that signal energy exists at points other than the immediate vicinity of the dipole itself. This leads to the idea that, if some of this signal energy from the surrounding area could be directed toward the antenna, the antenna would effectively have a greater signal input.

To increase this signal pickup effectively, the antenna employed is equipped with a "reflector" element. This is shown in Fig. 3-14 where a simple half-wave dipole is shown, and placed behind this dipole, in the same plane as the dipole, is the reflector element. This reflector conductor is usually of the same material as the dipole itself, and it should be slightly longer than the dipole. The reflector should be placed on the side of the receiving antenna *away* from the transmitting antenna from which the signal is to be received. This means that the desired signal will be approaching the antenna dipole in the direction indicated in Fig. 3-14. That part of the signal that passes the dipole and hits the reflector conductor will be reflected back to be picked up by the dipole.

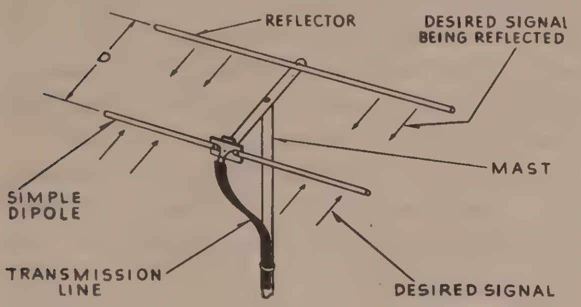


Fig. 3-14.—In order to increase the pickup, a reflector element is placed behind the dipole in the same plane at a distance *D*, that is between one-tenth and one-quarter of the wavelength for which the dipole is cut.

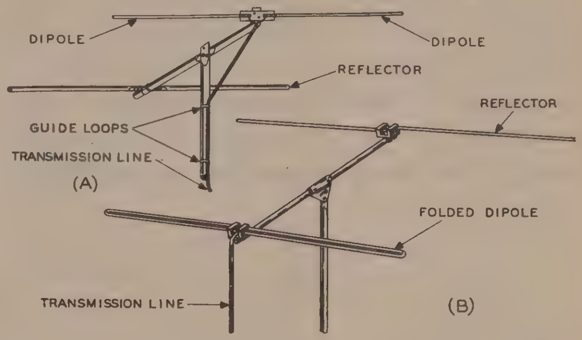
The distance *D* that the reflector is spaced from the dipole is a factor in the amount of increased signal pickup, and it is usually somewhere from one-tenth to one-quarter of a wavelength away from the feed-in dipole element. Most manufacturers specify the spacing in their service instructions accompanying the antenna. This is to make sure that the reflected signal picked up by the receiving dipole is *aiding* the signal directly picked up by the same dipole, so that the maximum possible total energy pickup is available.

The way in which the signal is increased can also be compared with transformer action. A certain amount of mutual impedance exists between the dipole and reflector, determined primarily by the distance separating the elements and the self-impedance of the elements. In brief, when the signal hits the reflector, a voltage is induced which causes a corresponding current to flow in the reflector. This reflector current, by analogous transformer action, induces a voltage into the lead-in dipole element, the magnitude depending upon the mutual impedance and the phase relation of the voltage depending upon the spacings between the elements, which is often one-quarter wavelength. Thus, it is seen how a receiver antenna with a reflector can have an effective increase in signal pickup over a half-wave antenna without a reflector.

Since a mutual impedance exists between the elements and since the signal pickup is increased, the effective input impedance is also somewhat increased. The exact value of increased input impedance depends upon the degree of coupling, the value of mutual impedance between the elements and, hence, the amount of reflected impedance. When used with a simple half-wave dipole or folded dipole, the reflector will increase the input impedance, so that, besides increasing the signal pickup, the reflector can change the input impedance to, perhaps, a better impedance match from the antenna to the receiver.

In Fig. 3-15 are illustrated some typical receiving antennas as used today. In part (A) the reflector is used with a simple half-wave dipole element and in part (B) the reflector element is used with a folded dipole element.

Since the reflector increases the signal pickup of the dipole element, it is customary to state how much this signal pickup is increased over the dipole element alone by so many decibels gain. Consequently, the reflector element is said to add so much decibel gain to the antenna.



(A) Courtesy Insuline Corp. of America
(B) Courtesy American Phenolic Corp.

Fig. 3-15.—Reflector elements are used not only with simple dipole antennas (A), but also with folded dipoles, as shown in (B).

When used with a reflector, half-wave antennas become unidirectional (that is, become more directional to signal pickup in one direction than any other) because there is very little signal pickup from the reflector side of the arrangement. In other words, the reflector element prevents those signals approaching the rear of the antenna from being picked up by the dipole.

This action of the reflector makes it extremely useful in preventing pickup of undesired signals by the dipole element. In many localities reflectors are not needed to increase the signal pickup because the signal strength in those areas is relatively strong. However, they are often used in such areas to prevent pickup of undesired signals.

Direct, Reflected, and Blocked Waves

The use of a reflector element is recommended for those localities where there are surrounding buildings, hilly terrain, or other signal obstruction mediums even if the signal pickup in the area is relatively strong. This is primarily to prevent the pickup of signals that may become reflected when hitting one of these obstructions. If such a reflected signal were to be picked up by the dipole element, there is the very likely possibility that "ghosts" on the picture tube would occur. Before discussing what these ghosts are and how they are formed, let us study the different paths by which the transmitted television signal may reach the receiver.

In Fig. 3-16 a number of *direct waves* are illustrated emanating from a television transmitting antenna. Two television receiving antennas are also shown in the drawing. Receiving antenna *A* does not receive any direct wave because the building on which it is located blocks all these waves. A direct wave that is obstructed by some medium and is thus unable to reach an antenna is termed a *blocked wave.* Such a wave is illustrated in Fig. 3-16. Of the other three direct waves illustrated in the drawing, one makes direct contact with receiving antenna *B* and the others strike two buildings at an angle and are reflected from these structures. Waves thus diverted are commonly known as *reflected waves.* In the drawing of Fig. 3-16 the reflected signals are picked up by the receiving antennas.

As long as the dipole element (that which has the transmission line attached) will intercept a television signal, no matter where it comes from, reception is possible. Thus reception by means of either direct or reflected waves or even a combination of these signals is possible. In Fig. 3-16 antenna *A* is seen to be receptive to only a reflected signal because reception of a direct wave is blocked by the building it is on. However receiving antenna *B* is receptive to both direct and reflected signals.

If the reflector element used on antenna *B* is so oriented that the *reflected wave* will be coming directly toward this reflector element (that is, the reflector element is broadside to the reflected signal), then this reflector element will greatly reduce pickup of this reflected signal by the dipole element. As mentioned, when an antenna receives both direct and reflected waves of the same transmitted frequency, then *ghosts* on the picture tube of the receiver will probably result. This is especially so when the signal strength of the reflected signal is appreciable compared to the direct signal.

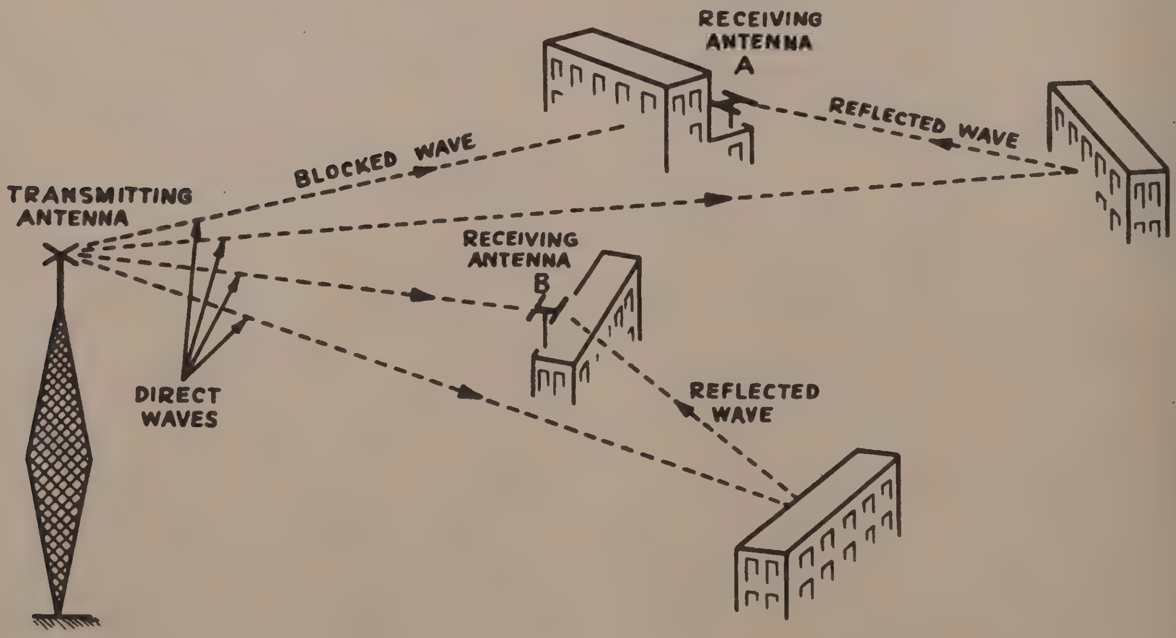


Fig. 3-16.—Receiving antenna *A* picks up only a reflected wave because the building it is on blocks the direct waves. Receiving antenna *B* is in the open and receives both a direct wave and a reflected wave, thus creating ghosts in the television receiver.

Ghosts or Multiple Images

We have used the word *ghosts* a few times in this chapter without stating exactly what they are. Ghosts need no introduction to many readers as they are quite familiar with this type of interference and know that it is essentially caused by reflected signals. Since this is still an important problem in the reception of good television pictures and also because many other readers do not fully understand the *different* methods by which ghosts are formed, a discussion of ghosts is in order.

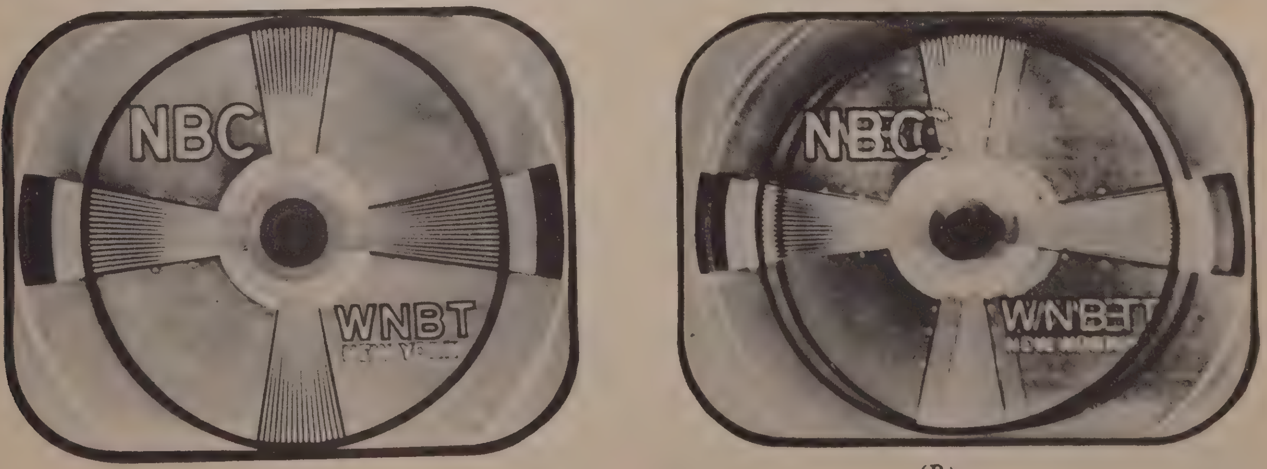
In brief, ghosts are multiple images formed on the screen of the picture tube. In Fig. 3-17 (A) a normal picture is illustrated and in Fig. 3-17 (B) is a picture showing multiple images or ghosts. In general ghosts are caused when the dipole element picks up two or more signals of the same transmitted frequency arriving at the receiving antenna at different instants of time. For instance in Fig. 3-16, the direct wave from the transmitting antenna takes a certain amount of time to reach receiving antenna *B* and the reflected wave (if picked up) takes a different amount of time to reach receiving antenna *B*. Due to the difference in time between the two signals, two separate images will appear on the screen of the picture tube, one for each signal received.

The direct signal is the strongest of the received signals and the reflected signals are the ones that produce the ghosts. Remember that each ghost is an exact image of the direct signal because all the signals causing images on the picture tube leave the transmitting antenna at the same time. Due to certain obstructions some signals are reflected and therefore take a longer path to the receiving antenna. This is readily evident in Fig. 3-16, where the path of the direct wave from the transmitting antenna to receiving antenna *B* is shorter in length than the path of the signal leaving the transmitting antenna, reflecting from a building, and finally reaching receiving antenna *B*. This difference in the length of the signal paths means that one signal will reach the receiving antenna before the other. The time difference in signal pickup produces ghosts or multiple images. The greater the number of reflected signals picked up, the greater the number of images on the screen.

In Fig. 3-17 (B) the ghosts appear to the right of the main picture and this form of multiple image reproduction is known as "lagging ghosts." In other words, the reflected signals (considered as the undesired signals when a direct signal is also picked up) are lagging in time with respect to the main or direct signal. When ghosts are due to reflections from surrounding buildings, the use of a reflector element and proper orientation of the receiving antenna is the best remedy for getting rid of such ghosts.

Ghosts may be caused in a television receiver by two other methods. In each case the ghosts are formed because the signals arrive at the input to the first tube of the receiver at different instants of time. The multiple images caused by one of these methods are known as "leading ghosts." In other words, the "undesired" signal arrives at the input to the first tube of the receiver before the "desired" signal. This situation is illustrated in Fig. 3-18.

Both signals involved are direct waves, but one signal is picked up by the antenna and the other signal is picked up directly at the input to the receiver. Since the path for signal *A* from the transmitting antenna to the receiver input is longer than that for signal *B*, there will be a time difference between the reception of each of these signals. Ghosts will, therefore, result, especially if signal *B* has appreciable sig-



(A) (B)

Courtesy R.C.A.

Fig. 3- 7.—A normal picture is shown at (A) and a picture with ghosts at (B).

nal strength when it reaches the input of the television receiver. At the input to the receiver, signal *A* will usually be the stronger because the antenna, which is a broadly tuned circuit, has a better response to signal *A* than the receiver itself to signal *B*. This means that the ghost or image caused by signal *B* will be displaced to the left of the main picture, and the ghosts in this instance are known as leading ghosts. In other words, the undesired signal (*B*) will be reproduced in the picture tube first and the desired and stronger signal will be lagging the image, or ghost, in this case; this also means that the ghost is leading the desired signal. This is the reverse of Fig. 3-17, where lagging ghosts are illustrated. The formation of leading ghosts is not as common as that of lagging ghosts. The best remedy in preventing these leading ghosts is to have the television receiver well shielded so that reception of signals directly through the receiver will be negligible.

Ghosts may also be caused by improper impedance matching of the transmission line to the receiver. Standing waves are consequently set up in the transmission line because reflections from the receiver will travel back to the antenna. When these signals hit the antenna proper, they act effectively as a reflected signal and enter the receiver retarded in time with respect to the original signal. In effect then, two signals of the same type but differing in time are accepted by the receiver, thus causing a ghost. This type of ghost will become more evident if the mismatched transmission line is quite long. In many instances, however, well-defined images are not readily evident, but a blurring of the pattern is the result.

Consequently we can see that besides causing loss in signal pickup, improper impedance matching of the transmission line to the receiver can also cause double images or blurring of the scene on the picture tube. A good impedance matching system using as short a transmission line as possible, is the best remedy for this type of trouble.

Maximum Voltage Input

It is well known that a maximum voltage input to the receiver is desired, but the feed-in to the simple half-wave dipole antenna or folded dipole is center driven and at this point the *current* is effectively at a *maximum*. How, then, can we conceive of a maximum voltage input to the first r-f tube? There are numerous way of explaining this, one of which follows.

In Fig. 3-19 a dipole antenna and input circuit to a television receiver is illustrated along with the current and voltage curves effective at the dipole. Since the center point of the antenna is at the loop of the current curve, a maximum amount of current will flow through the primary of the input transformer. This

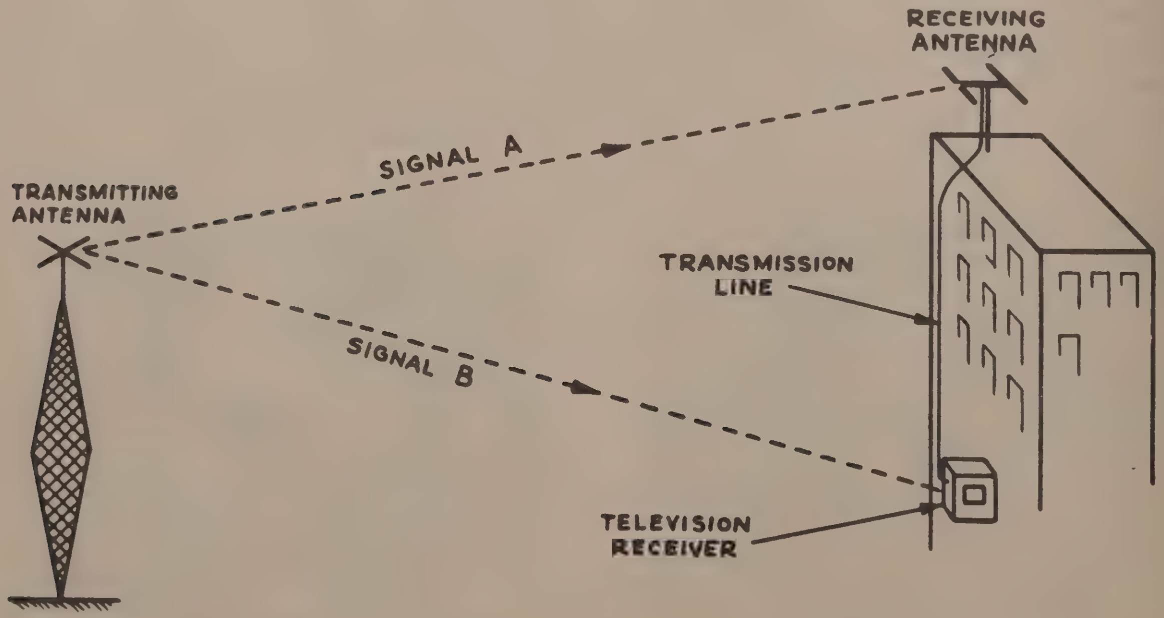


Fig. 3-18.—Two signals are received at the television receiver, one is picked up by the antenna and the other signal is picked up directly by the receiver.

maximum amount of current sets up a maximum magnetic field which causes a maximum voltage across the secondary due to induction. Since the parallel tuned secondary circuit is resonant at the frequency of operation, maximum voltage is developed at this desired frequency, which is applied to the grid of the input tube.

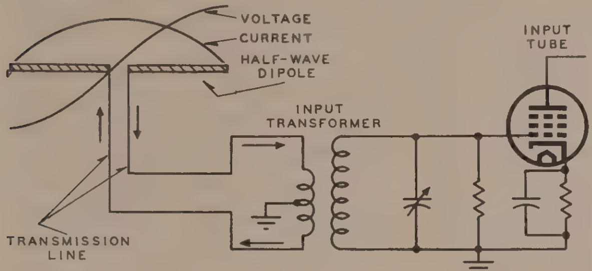


FIG. 3-19.—Half-wave dipole and input circuit to the r-f amplifier. The arrows indicate the current flow, which being a maximum in the transformer primary, induces a maximum voltage in the secondary at the frequency of operation.

Noise Reduction

In many television antenna systems, the primary of the input transformer has its center point grounded to reduce noise interference through the medium of the transmission line. This was illustrated in Fig. 3-19, but the left-hand side of this figure is redrawn in Fig. 3-20, to make this system of noise reduction somewhat clearer. Since the transmission line used may cover a greater area than the antenna itself, it has a tendency to pick up noise voltages, especially if the transmission line is quite long. To reduce this noise pickup and, hence, increase the signal-to-noise ratio at the input to the television receiver, the center tap of the primary is grounded. It reduces noise pickup in the following manner.

The noise signal, when it hits the transmission line, induces equal voltages in each lead of the transmission line which in turn produces noise currents that flow in the same direction in the transmission line as indicated in Fig. 3-20. By center tapping the primary of the input transformer to ground, this circuit becomes symmetrical, and the noise currents both flow toward this ground connection. This effectively makes the individual currents out of phase, and, since they are equal in magnitude, they produce magnetic fields which cancel each other and, hence, the total noise voltage induced in the secondary of the input transformer is virtually zero.

This reduction in noise pickup is only in reference to that picked up by the transmission line and not that noise picked up by the dipole itself. This latter noise, as well as the desired signal input, finds its way into the receiver.

Installation and Orientation of Antennas

It is beyond the scope of this book to discuss the installation and orientation procedures for all the different television antennas manufactured. The following notes, however, cover outdoor receiving antennas and apply to all types, unless otherwise specified.

One of the primary requisites in all installations is that the antenna system be mechanically secure. The first necessity is to make sure that high winds, rain, ice, or snow will not affect the mechanical secureness of the system. The reason for this is obvious, since it is desirable that the antenna require little attention once it is installed. Most types of antennas are supplied with the necessary mechanical supports to attach the system to some object such as a roof, chimney, or side of a house.

The antenna should be placed as high as possible and away from any interfering objects which might reflect the television signals. Consequently, such high installation points as roof tops are used. The antenna also should be mounted some distance away from any of the metallic objects commonly found on a roof top, such as water drains, pipes, and wires. It is quite as important to make sure that the transmission line used (especially if it is the twin-lead type) is also kept away from metallic objects. By placing the transmission line near such objects, noise and other types of undesired reflected signals can be picked up by the line.

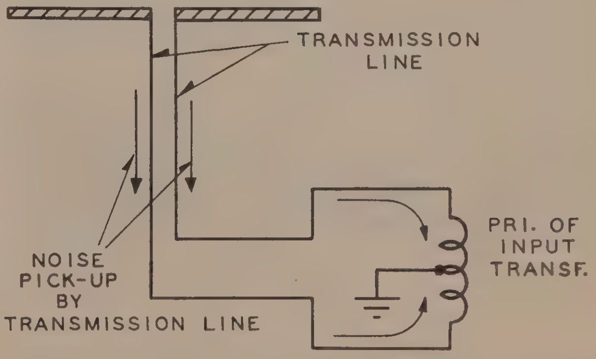


FIG. 3-20.—In order to reduce noise pickup through the medium of the transmission line, the center of the primary winding of the r-f transformer is grounded. This increases the signal-to-noise ratio at the input of the receiver.

Once the spot with the most freedom from metallic objects has been chosen, two remaining important factors must be taken into account: the orientation of the antenna proper and the correct installation of the transmission line from the antenna to the receiver. Because of the horizontal polarization of the radiated television signals, the radiating elements of the receiving antenna must be in a position to receive the maximum possible signal pickup. Most television receiving antennas, especially in installations containing dipole elements, are so oriented that the elements are horizontally situated.

In using antennas that are unidirectional or bidirectional, it is important to know something about the television transmitting stations within the coverage area of the receiver in question. If the field intensities of most of the transmitting stations within the vicinity of the receiver are about the same, not much difficulty will be encountered in orienting the antenna. If most of these stations are quite near each other as far as the receiver is concerned, the receiving antenna can be oriented simply by placing the pickup element *broadside* to the oncoming television signals. The receiving antenna thus serves its most useful purpose in trying to obtain equalized signal pickup as nearly as possible for all of the television stations in its area.

If the antenna is placed in a different position there may be a noticeable loss in signal pickup when the receiver is in operation. This is particularly true when the receiver is quite a distance from the television stations. However, as pointed out, in localities where the receiver is in the vicinity of strong local stations, variation in the position of the antenna may make little difference.

If the strengths of the various television signals in the locality of the receiver in question are quite different from each other, and if they approach the receiver from different directions, it is advisable to orient the antenna broadside to the signals that are weakest, in order to have adequate signal pickup on most stations.

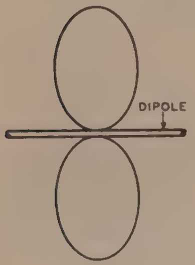


FIG. 3-21.—The ideal field pattern of a half-wave dipole. Note that each side of the pattern is identical.

The field pattern for a dipole, considered either in the receiving or transmitting sense, is a conventional figure-eight pattern, as shown in Fig. 3-21, with the points of maximum and minimum signal pickup 90° out of phase with each other. Therefore, when going from a minimum signal pickup to a maximum, the antenna should be rotated by 90°. This field pattern shows that the dipole is bidirectional.

The installation of the transmission line is as important an item as the antenna itself. The transmission lines most commonly used, as previously mentioned, are of two types: the twin-lead line (with plastic insulation) and the coaxial line. Both types can be supplied with different characteristic impedances. The importance of correct impedance matching for maximum energy transfer should be remembered, especially in localities where the received signals are weak. One advantage of the coaxial line is that it has low inherent signal loss per section and low noise pickup, as compared with the twin-lead type. However, the coaxial type is primarily used for low impedance matchings because high impedance coaxial lines are not available for these high-frequency impedance matching systems. The twin-lead type of transmission line at 300 ohms is the kind generally used for high impedance matching. Some of the factors that help determine the type of transmission line to use are the type of antenna (high or low impedance), input impedance of receiver, and length of transmission line needed.

In attaching either type line from the antenna to the receiver, the length of line used should be as short and as rigid as possible. The requirement for a short line arises from the fact that the longer the line, the more signal will be lost. The rigidity requirement is necessary to prevent the line from swaying with the wind. A number of things may happen if the line does sway in the wind. The line may be moved near some metallic object, it may be scraped constantly against something which will make for quick deteriorating of the insulator and then perhaps shorting of the line elements, or the line may even be snapped by a strong wind.

The mast and supports of the antenna should have some provision for stand-off insulators or guide loops, so that the portion of the transmission line leading from the antenna can use these insulators or guide loops as a means of preliminary support. These special attachments help maintain rigidity in the line and prevent any unnecessary swaying. Some manufacturers suggest that, when passing twin-lead line through these loops that are connected to metal masts, the line should be twisted two or three times to maintain electrical balance between each wire of the line and the metal mast and thus minimize noise pickup. However, these twists in the line should be made only in the vicinity of the metal mast.

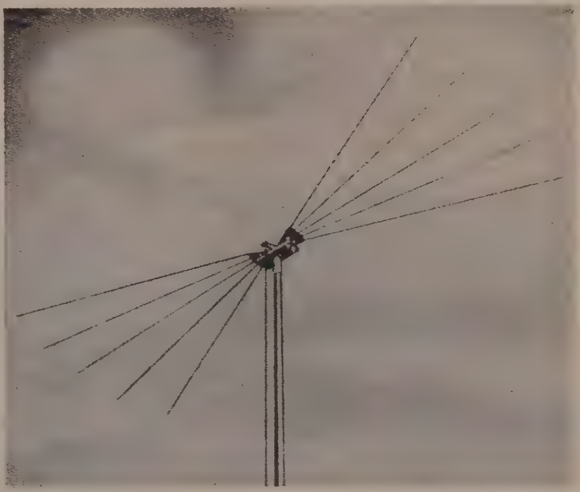
After passing through the insulators or guide loops, the line should be drawn tight until it is attached to the receiver. In doing this, the line may be pressed flat against only those types of objects that will not influence its transmission characteristics. Such things as wood, brick, and roof shingles (no metal involved) are commonly used. To secure this tightness in the line, special guide loops or other such attachments are often attached to some place in the vicinity where the transmission line is running. Twin-lead line may be fastened at a number of points by using small metal brads (that is, nails without any heads) through the center of the plastic insulator and into any material that will take the nail, provided the material is not metallic. The brads themselves have a negligible effect on the transmission of the signal along the line. However if nails with heads or similar objects are used, difficulties may arise since a wearing effect on the insulator may occur and the line may be short-circuited, or the characteristic impedance of the line at the point where the nails pass through it may be changed, causing a mismatch from the antenna to the receiver.

Other Types of Television Antennas

So far in the analysis of television receiving antennas we have dealt primarily with simple and folded dipole antennas, with and without reflectors, because they are the types that are most universally used. There are, however, numerous other types of antennas used for the reception of television signals. Some of them are no more than slight variations of the antennas already analyzed. Many of the design principles and the theory involved have already been discussed in general form as applicable to all antennas. These other types of antennas are usually designed for one or more of the following reasons: better signal pickup, omnidirectional signal response, and better broadband response.

Some of these antennas may be higher priced than those previously discussed. In other words, if more time and equipment is put into the design of television antennas, the price will be high compared with other antennas. Thus the television receiver buyer, or even the television manufacturer buying antennas, is often faced with limitations on how much he can economically afford to spend on antennas. Therefore, most antenna manufacturing companies refer to the electrical advantages or superiority of their product in respect to their price. They are trying to be as fair as possible to the public in establishing what they believe to be an honest account of what they have put into the antenna in the form of engineering design as well as material things.

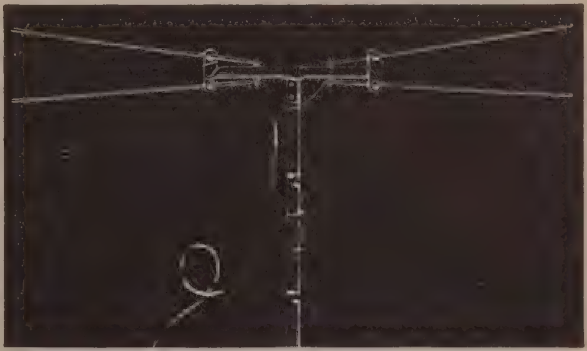
In Fig. 3-22 is illustrated an antenna known as the Di-Fan. This antenna offers broadband characteristics over the complete range of the two television bands and hence also includes the f-m broadcast band of 88 to 108 mc. The nominal impedance of this type of antenna is about 300 ohms and it is therefore suitable for use with a 300-ohm transmission line.



Courtesy Andrew Co.

FIG. 3-22.—A special type of antenna known as the Di-Fan. It offers broadband characteristics over the two television bands and the f-m band.

The antenna illustrated in Fig. 3-23 is known as the double-V. This type of antenna, when cut to the proper frequency in either television band, is said to give a fairly good frequency response over the entire band. When designed to be used with the low-fre-

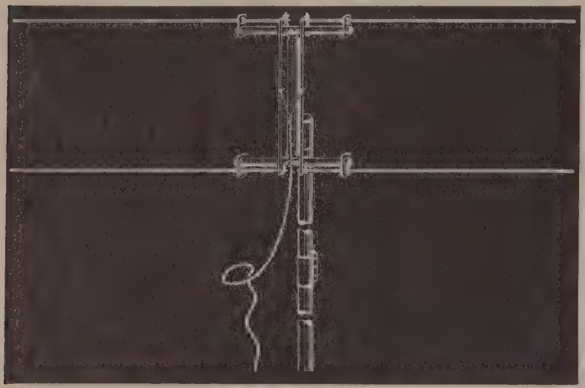


Courtesy Technical Appliance Corp.

FIG. 3-23.—The double-V antenna has a broadband characteristic.

quency television band it can also be used for the f-m band.

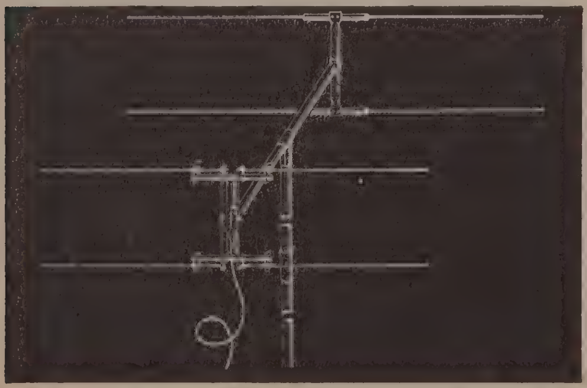
The stacked dipole, the so-called H-type antenna, is shown in Fig. 3-24. This type of antenna essentially consists of two center-fed dipoles mounted one above the other with their center points connected as shown. The transmission line is attached to the parts connecting these center points. This antenna, like a single dipole, is bidirectional. Mounting one dipole over the other allows for greater signal pickup in a broader horizontal plane. In other words, this system provides greater gain than the simple dipole. Since higher gain is obtainable, this antenna is recommended by the manufacturer to be used in localities where the signal strength is low.



Courtesy Technical Appliance Corp.

Fig. 3-24.—The stacked dipole, or H-type antenna, is bidirectional and provides greater gain than does a simple dipole.

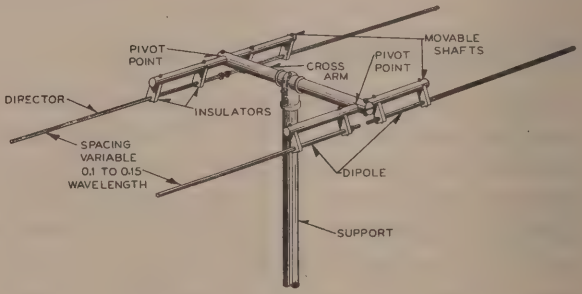
The stacked dipole of Fig. 3-24 is put to further use by employing two reflectors with the unit. This is shown in the antenna drawing of Fig. 3-25. This antenna is called the double-doublet by its manufacturer and essentially consists of one simple dipole and reflector arrangement mounted above the other. The dipole elements are, however, connected together at their center points as in the H-type antenna. The reflector elements are added to help prevent ghosts and to make the antenna unidirectional. The over-all gain of this antenna is somewhat greater than the H-type because of the addition of the reflector elements. It is recommended by the manufacturer that this antenna be used in areas where there are a number of reflected signals and also in noisy areas where the signal strength is low.



Courtesy Technical Appliance Corp.

Fig. 3-25.—The double-doublet antenna essentially consists of a simple dipole and reflector arrangement mounted above another one. The reflector elements are added to help prevent ghosts and to make the antenna unidirectional.

There are antenna systems appearing on the market that appear to be a dipole and reflector arrangement, but the two elements are spaced somewhat closer together than the usual dipole-reflector arrangement. In most cases these antenna systems use the element with the transmission line attached as a director and not a reflector. Director is the name given to an antenna parasitic element that is oriented in the direction of the signal to be received. The exact spacing between the dipole and director is determined by the physical makeup of the antenna system and surrounding territory. In Fig. 3-26 a dipole and director system is illustrated; the distance between the two elements is variable so that the maximum use of the director can be attained wherever the antenna is installed. The director element is made slightly smaller in length than the dipole element.

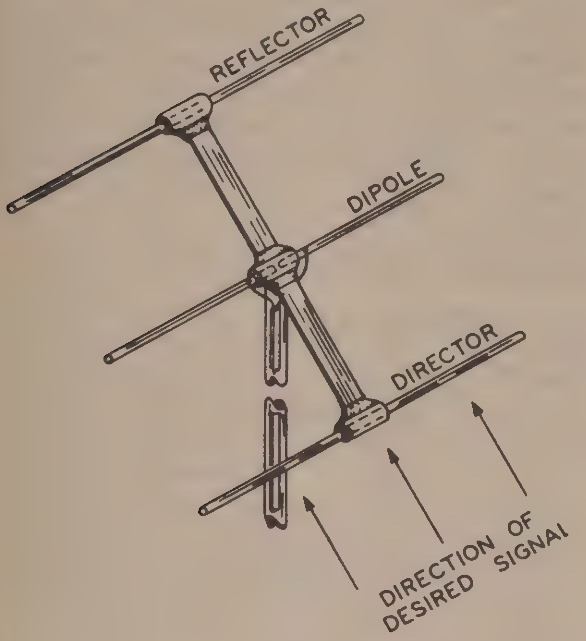


Courtesy Shur-Antenna-Mount, Inc.

Fig. 3-26.—The distance between the dipole and director is variable so that best use of the director can be attained when the antenna is installed.

The input impedance of many such systems is quite low, and special impedance matching sections very often must be used in conjunction with the transmission line so that the maximum amount of energy transfer from the antenna to the receiver can be attained. Most manufacturers incorporate the necessary equipment for correct impedance matching, or include information as to how it may be accomplished.

The signal pickup can be further increased by an antenna system employing both director and reflector elements in conjunction with a dipole. A drawing of such a system is illustrated in Fig. 3-27, where the dipole is the center element. Although not readily evident in this drawing, the director element is slightly shorter than the dipole and the reflector element slightly longer than the dipole. The director is placed in the direction of the television signal. This type of antenna system, in addition to providing increased signal pickup, increases the directivity as compared to the dipole reflector and dipole-director antennas. That is, this antenna is unidirectional and the reflector element also helps eliminate ghosts.

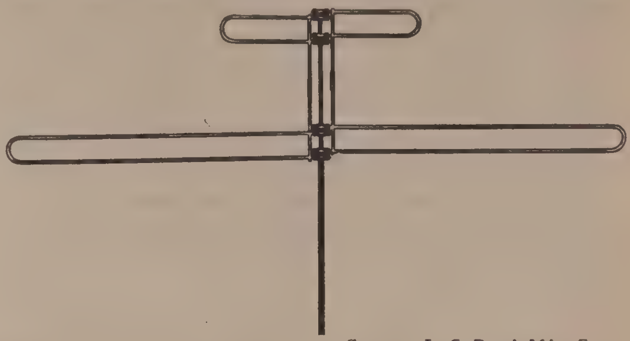


Courtesy Consolidated Television Corp.

Fig. 3-27.—The director is placed in the path of the incoming signal and the reflector is placed behind the dipole. Increased signal pickup and great directivity are provided by the use of this system.

There are a number of television antennas on the market today that are so designed that they contain two separate dipole elements, one for each television band. The element designed for the high-frequency band (channels 7-13) is smaller than the element designed for the low-frequency band (channels 1-6). In most cases the elements are cut for the center frequency of the band in question.

Fig. 3-28 illustrates one such antenna. Upon first inspection, this antenna appears as though it consists of two folded dipoles. This is not the case. Each element is a simple dipole that is folded back on itself in a U shape and shorted at the ends. The *Q* of each element is made smaller by this method, thereby making the antenna frequency response broader. This special shaping also adds to the rigidity of the antenna system. The longer antenna is for the low-frequency television band and the shorter antenna for the high-frequency band. To prevent interaction between antenna elements, they are separated by special sections called "dividers." These dividers are the black portions of the vertical rods connecting the two antenna elements as seen in Fig. 3-28. The antenna elements and dividers are so designed that by connecting the transmission line above the dividers, good reception on both television bands will be achieved.



Courtesy L. S. Brach Mfg. Corp.

Fig. 3-28.—Each element is a simple dipole, folded back on itself and shorted at the ends. The longer element is for the low-frequency television band and the shorter element for the high-frequency band.

All the antennas discussed so far were either bidirectional or unidirectional. There are some antennas however, that are *omnidirectional* in that they will respond to signals from all directions. Such an

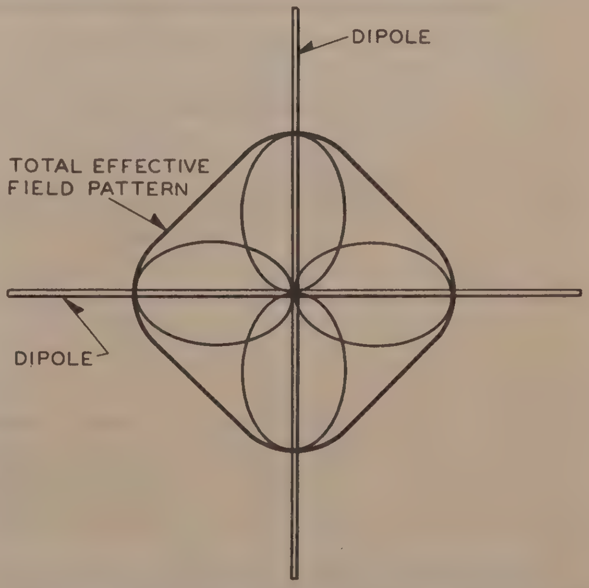
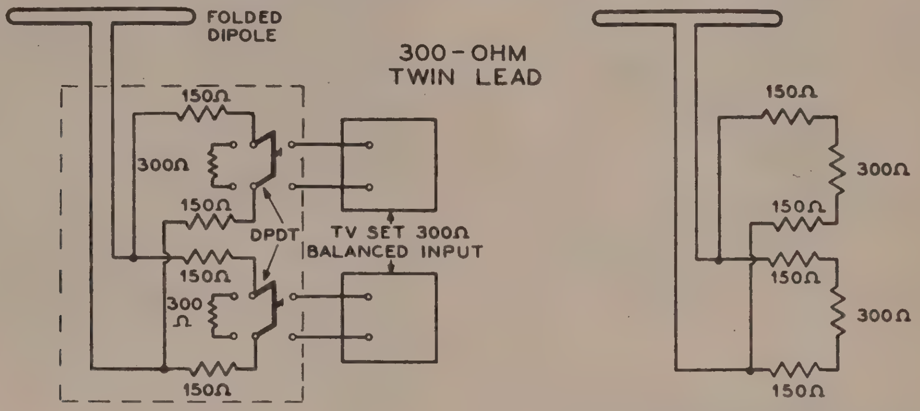


Fig. 3-29.—The total effective field pattern of two crossed dipoles is virtually circular in shape.

antenna is, of course, only suitable in regions where there is very little chance for reflected signals and thus little possibility of ghosts. Omnidirectional antennas have been used a great deal with f-m receivers with very good results. These antennas essentially consist of two dipole elements both of the same size, at right angles to each other. Their lengths are cut to the center of either band, or whatever other frequency may be desired. Their center-fed points are usually individually connected to transmission lines, which in turn are combined in a single line for whatever the impedance desired.

Two crossed dipoles are drawn in Fig. 3-29 in conjunction with the field patterns of each individual dipole. The *total effective field pattern* is virtually circular in shape due to the addition of the field patterns of the individual dipoles making up the antenna. Such an antenna is definitely suitable in areas where there are a number of different television signals coming from different directions. In such areas, special orientation of the antenna to the weakest signal, as is usual with bi- or unidirectional antennas, is not necessary because the antenna will respond equally to all signals.

TWO OR MORE RECEIVERS ON ONE ANTENNA

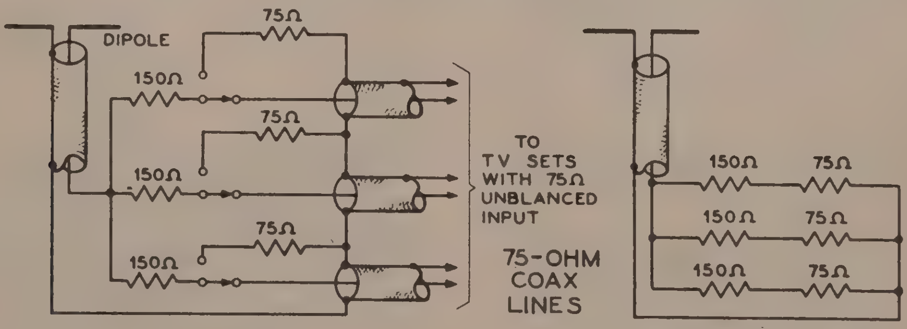

FOLDED DIPOLE
300-OHM TWIN LEAD
150Ω
300Ω
150Ω
DPDT
TV SET 300Ω BALANCED INPUT
150Ω
300 Ω
150Ω
150Ω
300Ω
150Ω
150Ω
300Ω
150Ω

(A)

(B) EQUIVALENT CIRCUIT

SERIES RESISTANCE FOR

TWO RECEIVERS ——→ 150Ω
THREE RECEIVERS ——→ 300Ω
FOUR RECEIVERS ——→ 450Ω


DIPOLE
75Ω
150Ω
75Ω
150Ω
75Ω
150Ω
TO TV SETS WITH 75Ω UNBLANCED INPUT
75-OHM COAX LINES
150Ω
75Ω
150Ω
75Ω
150Ω
75Ω

(A)

(B) EQUIVALENT CIRCUIT

SERIES RESISTANCE FOR
TWO RECEIVERS ——→ 75Ω
THREE RECEIVERS ——→ 150Ω
FOUR RECEIVERS ——→ 225Ω

You must know the relative signal strength in your area so you can tell if 2, 3, or 4 receivers can be connected to one antenna, as shown above. For two sets, the available signal voltage is cut 2 to 1; for three, it is cut 3 to 1, and for four sets, it is 4 to 1. Use carbon resistors, as wire wound may unbalance the system. Place switches and resistors in a control box. All leads must be short and direct. If a set is removed from the system, throw switch to dummy load.

CHAPTER 4

R-F AMPLIFIER, OSCILLATOR, AND MIXER CIRCUITS

By Henry Chanes

The front end of a television receiver includes the r-f amplifier, if any, the oscillator, and the mixer. Usually these circuits are assembled together as a tuner which supplies the i-f inputs for the video and sound i-f amplifiers. The functions of the r-f circuits in a television receiver are similar to the functions of the corresponding circuits in the conventional superheterodyne receiver. The r-f amplifier selects the desired signal, amplifies it, and rejects all undesired signals. This amplified signal is combined in the mixer tube with a locally generated signal of a frequency that differs from the incoming signal by a fixed amount. When the two signals are mixed, a beat frequency is produced in the output of the mixer. This beat frequency, which is equal to the difference between the two signals, is known as the intermediate frequency (or i.f.) and is fed tc the input of the i-f amplifiers. This, essentially, is the function of the front end in the television receiver, and so far it is the same as the front end of an ordinary sound superheterodyne receiver. However, because of the higher frequencies and bandwidths required by television, the front end design of a television receiver becomes somewhat more complicated.

The R-F Amplifier

As in the case of sound receivers, not all television receivers employ an r-f amplifier before the mixer. In the receivers that do not, the signal from the antenna is fed directly to the input circuit of the mixer. There are several advantages to be gained by the use of an r-f amplifier.

1. The selectivity of the receiver will be increased, resulting in a greater rejection of image frequencies.

2. The signal-to-noise ratio is better when an r-f amplifier is used before the mixer stage. This is an important factor in the performance of a television receiver and will be discussed in detail later on.

3. The r-f amplifier decreases the amount of signal that will leak through to the antenna from the local oscillator. This signal might be radiated from the antenna and cause interference with other television receivers in the same neighborhood. This is especially objectionable in large cities where there are as many as 40 or 50 apartments in a single building. If each apartment contained a television receiver, the problem of local oscillation could indeed be serious.

One of the objections to the use of an r-f amplifier is that it increases the number of tuned circuits. If a switch is used, this means adding a large number of switching elements and the components of the tuned circuits. If continuous tuning is used, the problem of tracking is complicated. These complications result in a higher-priced receiver. There is also the fact that the gain of an r-f amplifier in a television receiver is comparatively low and that much higher gain can be achieved if an extra stage of intermediate amplification is employed.

Generally the sound and picture carriers are not separated before the i.f.'s are obtained. Therefore it is necessary for the r-f amplifier and the mixer tuned circuits to have a bandwidth sufficient to pass the entire television channel of 6 mc. (The bandwidth of a television channel is discussed in chapter 2 and illustrated in Fig. 2-4.) This bandwidth is not too difficult to obtain at the frequencies used for television channels. Even at the lowest channel, which is from 44-50 mc, the 6-mc bandwidth is not too great a percentage of the carrier frequency. The tuned circuits themselves are usually wide enough to take care of this frequency range. If not, resistors can be put across the tuned circuits to load them down and increase the bandwidth.

Another very important consideration in the design of the r-f amplifier is the signal-to-noise ratio. At the first stage in the receiver, the signal is at a very low level, and any noise developed in this stage will be amplified along with the signal; and if the noise is a large enough percentage of the signal, it may mask out the picture entirely. For a good picture, the noise level should be about 40 db below the signal level. By noise we do not mean the noise picked up by the antenna or transmission line, that is, atmospherics and man-made noise; that noise should be taken care of by a suitable antenna and transmission line. We are now considering the noise actually generated in the r-f amplifier or, if an r-f amplifier is not used, the noise generated in the mixer. This noise is caused by both the thermal and shot effects. Noise due to thermal agitation is due to the fact that the electrons in a conductor are always in motion. When no current is flowing through the conductor, this motion is random and the amount of motion will increase with the temperature of the conductor. When current is flowing, the average flow of electrons will be in one direction, but a random component of motion due to thermal agitation will still exist. This noise will increase as the temperature, impedance, or bandwidth is increased.

The other component of the noise generated in the r-f amplifier is due to shot effect. When the electrons leave the cathode in a vacuum tube, they do not flow in a smooth stream to the plate but leave in small groups. These groups cause pulses of current in the plate circuit that constitute a random noise. The strength of this noise signal increases with an increase in the impedance in the plate circuit, the current, or the bandwidth. These two sources of noise in the r-f amplifier, rather than the amount of gain that can be achieved in a television receiver, are the limiting factors on the amount of signal necessary at the input of the receiver in order to produce a good picture. Taking these factors into account, and assuming a minimum amount of man-made noise, a signal strength of 500 microvolts at the input to the receiver should provide satisfactory performance. In noisy locations where the interference is great, larger signals than this may be necessary.

Input Circuits

The antenna, or the transmission line if one is used, must be matched to the input of the r-f amplifier, or the mixer, if the receiver does not have an r-f amplifier. If the line is not properly matched, reflections will occur that will interfere with the signal and result in a blurred picture. The simplest type of coupling circuit consists of a transformer with a tuned secondary connected to the grid of the tube. The primary, which is untuned, is connected to the transmission line. This is shown in Fig. 4-1. The inductance of the primary is usually very small compared with that of the secondary, and its effect on the tuned circuit will be negligible. The resonant frequency of this input circuit will be determined by the inductance of the secondary L and the shunting capacitance C. This capacitance is made up of the input capacitance of the tube, stray capacitance due to the wiring, and the minimum capacitance of a tuning capacitor if one is used. As in any resonant circuit, tuning may be done by varying either L or C. Either one or both may be a preset control which is adjusted for a particular station and switched into the circuit when desired, or one may be variable for continuous tuning. The resonant frequency of this circuit is given by the expression:

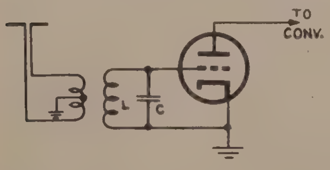


Fig. 4-1.—The simplest type of transformer coupling is the single-tuned circuit.

$$f_r = \frac{1}{2\pi\sqrt{LC}} \qquad \text{Eq. 4-1}$$

where:

f_r = resonant frequency, in cycles
L = inductance in henrys
C = capacitance in farads
π = constant = 3.14 approximately

Usually the losses in the secondary will be great enough to give this circuit sufficient bandwidth; if not, resistance can be shunted across the secondary winding. When continuous tuning is employed, several problems are encountered because of the range of frequencies that have to be covered. The highest frequency in the television band is almost five times the lowest frequency. If C is made a variable, the minimum value of a capacitor with a suitable range will cause a decrease in voltage gain, and in order to keep a constant gain over the band, the mutual inductance between the primary and the secondary will have to be changed as the circuit is tuned to different frequencies. If L is the variable, there is the problem of getting sufficient change in inductance to cover the frequency range. These problems have been solved by the use of special circuits that will be discussed.

If the input is balanced, as in Fig. 4-1, the center tap of the primary winding is grounded. An interfering signal would tend to induce equal voltages in the two wires of a transmission line, since they are closely spaced. This would cause currents that are equal but opposite in polarity to flow in each half of the primary. The effect of the two would cancel and the interfering signal would not appear on the secondary. The signal due to the antenna would not be canceled because the antenna produces a *voltage* on the transmission line that is balanced to ground. That is, the instantaneous voltages on the two wires of the transmission line are always of opposite polarity. The currents in the primary of the input transformer due to these voltages would add up and produce a voltage in the secondary. If a coaxial type of transmission line is used, the input is not balanced and one end of the primary winding is grounded. In this case, interference pickup by the transmission line is prevented by the shielding provided by a coaxial cable. In this respect, the coaxial cable is more effective than the open-wire line with a balanced input but is also more expensive.

The gain of the circuit shown in Fig. 4-1 may be improved by tuning the primary as well as the secondary of the transformer. This circuit is shown in Fig. 4-2. Both circuits are tuned to the same resonant frequency which again may be determined by Eq. 4-1. While this circuit has greater gain than the single-tuned circuit, it has the disadvantage of having another tuned circuit, which complicates the switching or tuning circuits, whichever is used. To obtain a voltage gain in the input transformer, the secondary inductance should be large with a correspondingly low value of shunting capacitance. The minimum value of C_s is limited by the input capacitance of the tube, making tubes with interelectrode capacitances desirable. In order to keep the ratio of inductance to capacitance as large as possible, C_s is usually the input capacitance of the tube and strays due to wiring, and the tuning is accomplished by varying the inductance. This may be done by having taps on the coil, by switching in different coils, or by changing the position of an iron core in the coil.

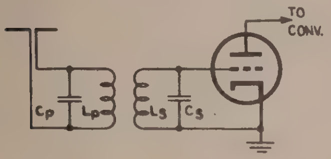


Fig. 4-2.—The gain of this double-tuned circuit is improved over that of Fig. 4-1.

Frequency Converters

The output of the r-f amplifier is fed to the mixer. If no r-f amplifier is used, the antenna, or transmission line, is connected to the mixer through a suitable input circuit. This input circuit may be of the same type that is used for the r-f amplifier. In Fig. 4-3 we have a typical mixer circuit. For the sake of clarity, loading resistors across the tuned circuits have been omitted, and the mixer tube has been shown as a triode although it usually is a pentode. The r-f amplifier is coupled to the mixer through a single-tuned circuit consisting of C_1 and L_1. This circuit is tuned to the same frequency as the input circuit of the r-f amplifier, and the switching or tuning considerations mentioned will also be true of this circuit. The signal is then coupled to the mixer tube by the coupling capacitor C_c. If the circuit were double-tuned, the secondary of the transformer would be connected to the mixer grid. in this particular circuit the output of the local oscillator is coupled to the same grid by the capacitor C_o. In some circuits inductive coupling is used between the oscillator and the mixer. The two signals are mixed to produce a third frequency which is the difference of the two. The primary and the secondary of the output transformer, C_2L_2 and C_3L_3 respectively, are both tuned to the difference frequency. The output is the intermediate frequency or i.f. and is fed to the i-f amplifiers.

The local oscillator is operated at a frequency above the picture and sound carriers. Since the sound carrier is always 4.5 mc above the picture carrier, the sound i.f. will be 4.5 mc below the picture i.f. Let us consider the second channel 54 to 60 mc, as an example. The picture carrier is 55.25 mc and the sound carrier is 59.75 mc. In a typical receiver, the local oscillator is set to operate at 82 mc when tuned to this channel. This produces a picture i.f. of 26.75 mc and a sound i.f. of 22.25 mc. It is advantageous to have the picture i.f. at the higher frequency because of the larger bandwidth required for the video signal. At the higher frequency, this bandwidth is a smaller percentage of the carrier frequency and requires less loading of the tuned circuit. This means a higher value shunting resistor and in turn, a higher gain.

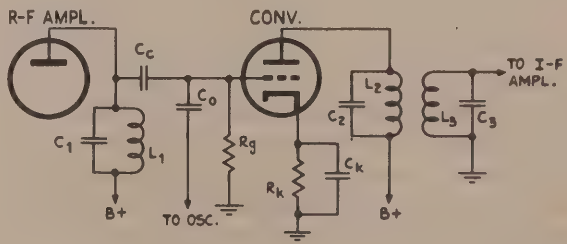


Fig. 4-3.—Simplified schematic of a converter (mixer) stage.

In almost all television receivers, separate tubes are used for the mixer and local oscillator, as the combination tubes available at present are not satisfactory. These tubes do not have a high enough conversion gain and do not oscillate readily at the high frequencies required for television.

High-Frequency Oscillators

The oscillator used in a television receiver must be able to supply sufficient signal over the entire range of frequencies and has to be stable with respect to voltage and temperature. To cover the entire television band, the oscillator will have to work over a range of frequencies from about 70 to 240 mc. The oscillator tuning accuracy is an important consideration in television receivers, since the oscillator determines the i.f. for both the picture and sound channel. At the same time that the picture i.f. is at the correct frequency, the audio i.f. has to be placed in the center of the response band for proper reception of the sound. Generally, the audio i-f amplifiers will have a bandwidth of about 250 kilocycles. The f-m sound signal only swings 25 kc each side of the carrier, and this bandwidth is not necessary just to pass the sound signal. However, by using a larger bandwidth than necessary, it is possible to build an oscillator that will keep the audio i.f. within the pass band. In a receiver of this type, the audio i.f. will be able to drift 100 kc above or below the center frequency and still be passed by the sound i.f. amplifier. This means an allowable drift of the oscillator of 100 kc in 240 mc at the highest oscillator frequency, or about 0.04 percent. This requires a very stable oscillator and also an accurate method of adjustment.

High-frequency oscillators are more subject to drift from the original frequency of operation, due to such factors as heat, humidity, and changes in the B-supply voltage on the elements of the oscillator tube, than are low-frequency oscillators. The oscillator drift caused by heat and humidity is a slow process, wherein the elements of the oscillator tube and the circuit components change their inductance and capacitance values in accordance with the slow changes in temperature and humidity surrounding the complete oscillator circuit.

The effect of humidity on oscillator drift is not readily apparent in locations where the atmosphere is arid and the humidity does not undergo much change; however, it is definitely apparent in places having a humid atmosphere which is constantly changing. The effect of humidity is such that a certain amount of moisture condenses on the coil and capacitor of the oscillator tank circuit and causes a change in the dielectric surrounding them. As far as a variable air capacitor is concerned, the moisture collects on the plates and changes the dielectric between them. Since the dielectric is a determining factor in the amount of capacitance, it is evident that the collection of moisture on the plates changes the value of capacitance, which in turn changes the frequency of operation of the oscillator. The moisture that collects on the coil of the oscillator tuned circuit changes the dielectric properties of the surrounding medium, which in turn changes the distributed capacitance of the coil, resulting in oscillator frequency drift.

The bad effects of humidity are usually controlled by adding some unit that will produce a constant high temperature in the vicinity of the oscillator, thereby keeping the circuit dry. Another method is to prevent moisture by placing the oscillator tank circuit in some form of impregnated can. Still a third method is to coat the active coil and capacitor elements that may be affected by humidity with some moistureproof material.

Frequency changes due to humidity do not occur very often compared with those due to heating effects. Changes in temperature surrounding the oscillator circuit will cause the inductive and capacitive tank circuit components to change in value, thereby causing the oscillator to drift in frequency. It should be remembered that we are dealing with very high frequencies (the oscillator in television receivers operating in the vicinity of 70-240 mc) and that changes in the shape of an inductance or capacitance, although minute, will cause a definite change in oscillator frequency.

Increases in temperature cause the windings of the coil and plates of the capacitor to expand, thereby inherently increasing the inductive and capacitive values of the oscillator tank circuit. The amount of expansion is determined by the material from which the coil and capacitor are made, and a constant called the *temperature coefficient* is a ready means of determining how much the component will expand. A low temperature coefficient means that the component will have a small amount of expansion and, therefore, contribute little to oscillator frequency instability. Low-temperature coefficient coils and capacitors are desired in high-frequency oscillator circuits. Since it is more difficult to obtain a variable capacitor with a low-temperature coefficient than a variable inductor with this feature, permeability tuning is preferable to capacitor tuning wherever temperature changes are evident in the oscillator circuit.

Oscillator instability due to temperature changes is chiefly controlled by the insertion of a negative temperature coefficient capacitor which is placed across the oscillator tuned circuit. When the temperature increases, the capacitance of this negative coefficient capacitor *decreases*, offsetting the increase in the values of the oscillator tank circuit components and thereby maintaining stability of the oscillator against temperature changes. The chief drift in the oscillator is during the initial warm-up period of the receiver.

Another cause of oscillator instability is change in the B-supply voltage on the oscillator tube. A change in plate voltage will cause the transconductance and plate resistance of the tube to change, which causes drift in the oscillator frequency. This change in supply voltage is usually due to line voltage variations. A regulated power supply can be used to stabilize the plate voltage, but this may not be necessary if the oscillator is carefully designed so that the frequency changes very little with moderate changes in supply voltage. Different types of oscillators have different degrees of stability with respect to changes in plate voltage.

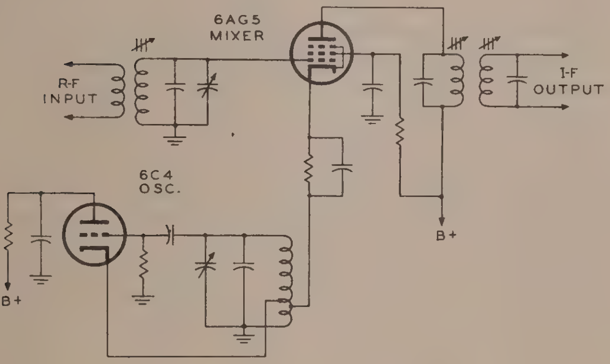


Fig. 4-4.—A typical oscillator-mixer circuit as used in television receivers.

A typical oscillator and mixer circuit that may be used in television receivers is shown in Fig. 4-4. The 6AG5 mixer tube and 6C4 oscillator are miniature type tubes especially designed for high-frequency work. Their input, output, and interelectrode capacitances are smaller than those of regular size tubes, making them readily adaptable to high-frequency work. The separate oscillator is a conventional Hartley circuit in which the coil is tapped and connected to the cathode. The 6AG5 mixer tube is a pentode with the control grid used for the r-f voltage input and the cathode circuit for the injection of the oscillator voltage. The method used to inject the oscillator voltage into the mixer tube in Fig. 4-4 is somewhat different from the coupling method used in the circuit shown in Fig. 4-3. Some of the oscillator voltage is tapped off part of the oscillator tank coil and fed directly into the cathode circuit of the 6AG5 mixer tube through the cathode's resistor-capacitor bias combination. Within this 6AG5 tube the r-f signal and oscillator signal are mixed together and form the i.f. which is selected by the i-f output tuned transformer circuit.

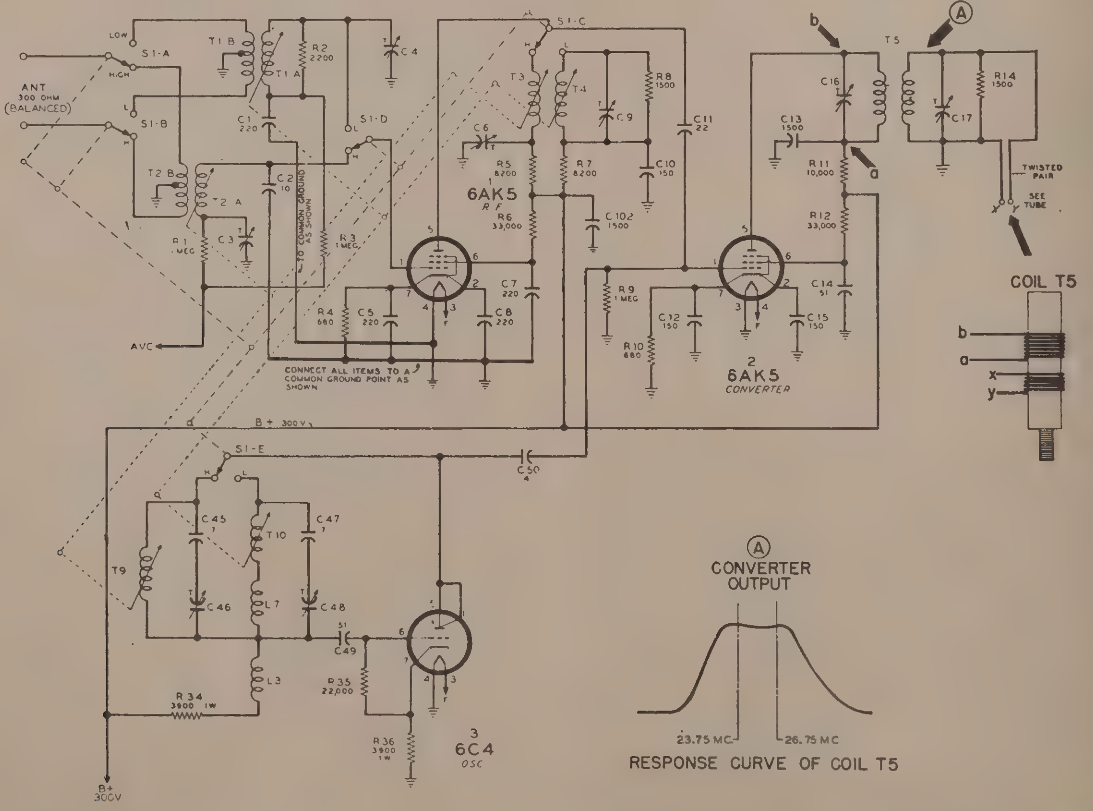
Belmont Model 21A21

The front end of the Belmont model 21A21 television receiver illustrates several of the features that we have discussed. The television channels are covered in two bands, the low band covering channels 1 to 6 and the high band covering channels 7 to 10. Separate tuned circuits for the high and low bands are used in the r-f amplifier input, the mixer input, and in the oscillator tank. These tuned circuits are switched by the switch *S*1. Within the band, the receiver is tuned to the different television channels by varying the inductance of the tuned circuits. This is done by changing the position of a powdered iron core in each coil. When the core, or slug, is entirely within the coil, the inductance will be higher than when the core is partly removed from the inside of the coil. The cores are all connected together mechanically and are adjusted simultaneously by a single tuning control from the front panel. This control is the tuning control that tunes the receiver to the desired station.

Let us consider this circuit in the high-band position, as shown in Fig. 4-5. The transmission line, which should be a 300-ohm balanced line, is connected to the transformer primary, *T*2-*B*. The secondary winding *T*2-*A* has an adjustable core, as mentioned above, and tunes with the capacitor *C*2 to the desired channel. The trimmer *C*3 enables this circuit to be aligned for proper tracking. The bandwidth of this tuned circuit is sufficient without a loading resistor, because the coil *T*2-*B* covers the high band of the frequency range, and the 6-mc bandwidth required is not too great a percentage of the carrier frequency. However, a loading resistor is necessary for the low-frequency band. This is resistor *R*2 which is across the low-band coil *T*1-*A*. The low end of the coil *T*2-*A* is connected to the avc circuit. The action of this is similar to the avc circuits used in sound receivers. The gain of the r-f amplifier is varied by changing the bias on the grid. The amount of avc voltage depends on the strength of the signal that is received. Besides the r-f amplifier, the first picture i-f amplifier, and the first sound i-f amplifier are controlled by the avc.

In the plate circuit of the 6AK5 r-f amplifier, the coil *T*3 is tuned to resonate with the output capacitance of the 6AK5 at the same frequency as the input circuit. The trimmer *C*6 has a similar function as the trimmer *C*3 mentioned above. The signal from this tuned circuit is coupled to the grid of the 6AK5 mixer tube by the capacitor *C*11. At the same time, the signal from the local oscillator is fed to the grid of the converter by the coupling capacitor *C*50. The local oscillator operates at a frequency 26.75 mc above the picture carrier frequency, therefore a picture i.f. of 26.75 mc will be produced in the mixer. Since the sound carrier frequency is 4.5 mc above the picture carrier frequency, a sound i.f. of 22.25 mc will also be produced in the mixer. The transformer in the output of the mixer, *T*5, is double-tuned to a center frequency of 25.25, is overcoupled, and has the secondary loaded by the resistor *R*14 in order to provide a response wide enough to pass the video and sound i.f.'s. This is shown at (A) in Fig. 4-5. The sound i-f frequency of 22.25 mc will come just within the pass-band of the coil *T*5. The picture i.f. of 26.75 mc does not fall at the center of the pass-band, but nearer to the edge of the response characteristic. This is because of the vestigial sideband type of transmission that is used for the picture channel. This is discussed in chapter 2.

The output of the converter is connected to the first video i-f amplifier. The audio i.f. is taken off the output of the first video i-f stage. The oscillator circuit is a modified Colpitts type and uses a 6C4 tube. At the high frequencies used in a television receiver, the interelectrode capacities are sufficient to be used as the feedback circuit in the oscillator. In this circuit the cathode is effectively connected to the junction of two capacitances that consist of the cathode-to-grid and cathode-to-plate interelectrode capacitance. On the high-band position, the frequency of the oscillator is determined by the tuned circuit made up of the coil *T*9 and the capacitors *C*45 and *C*46. The inductance of the coil *T*9 is varied to provide tuning to the television channel desired and is coupled to the other variable inductance coils. The trimmer *C*46 serves as an alignment adjustment to secure proper tracking of the oscillator. The choke coil *L*3 and the resistor *R*34 keep the oscillation out of the B supply and also serve to keep the grid from being grounded



LOW
S1-A
HIGH
ANT
300 OHM
(BALANCED)
L
S1-B
H
T1 B
T1 A
R2
2200
C4
C1
220
S1-D
T2 B
T2 A
C2
10
TO COMMON GROUND AS SHOWN
R1
1 MEG
C3
R3
1 MEG
AVC
R4
680
C5
220
CONNECT ALL ITEMS TO A COMMON GROUND POINT AS SHOWN
S1-C
T3
T4
C6
R5
8200
1
6AK5
R F
R6
33,000
C7
220
C8
220
R7
8200
C9
R8
1500
C10
150
C102
1500
C11
22
R9
1 MEG
R10
680
C12
150
2
6AK5
CONVERTER
b
A
T5
C16
C13
1500
R11
10,000
a
R12
33,000
C14
51
C15
150
R14
1500
C17
TWISTED PAIR
SEE TUBE
x
y
COIL T5
b
a
x
y
B + 300 V
S1-E
C50
4
C45
7
C47
7
T10
T9
C46
L7
C48
C49
51
L3
R34
3900 1W
R35
22,000
R36
3900
1W
3
6C4
OSC
B+
300V
A
CONVERTER OUTPUT
23.75 MC
26.75 MC
RESPONSE CURVE OF COIL T5

Courtesy of Belmont Radio Corp.

Fig. 4-5.—The front end of the Belmont model 21A21 television receiver. The switches are shown in the high-band position.

through the B supply. The cathode circuit also contains an isolating resistor *R*36 which helps maintain oscillation throughout the frequency range required by placing the cathode above ground.

General Electric Model 802

The local oscillator in the GE model 802 television receiver is used for the 13 television channels, the f-m band, and the broadcast band. The r-f amplifier and mixer used for the television band is also used for the f-m band. The schematic in Fig. 4-6 shows the front end of this receiver when the receiver is operating on the television band. For the sake of clarity, all switching circuits have been eliminated. Television channels are selected by a switch which changes the coils L_K in the cathode circuit of the r-f amplifier, L_P and L_S in the coupling circuit between the r-f amplifier and the mixer, and the coil L_O in the oscillator circuit. A 6AU6 tube is used as the r-f amplifier in a grounded-grid amplifier circuit. In this type of circuit, the grid of the amplifier is connected to ground, and the input is connected to the cathode circuit instead of to the grid as in the conventional amplifier. The output load is still in the plate circuit. There are several advantages to the use of a grounded-grid amplifier at the high frequencies utilized for the reception of television signals. The grid, being grounded, acts as a shield between the cathode and the plate. This prevents coupling between the input and output circuits due to the interelectrode capacitances that otherwise might cause oscillation unless the amplifier were neutralized. Neutralizing an amplifier over the wide frequency range encountered in television would be difficult and the grounded-grid amplifier eliminates this necessity. Another important advantage is that the output capacitance is reduced. The output load is across the plate to grid circuit rather than the plate to cathode circuit. The interelectrode capacitance between the plate and grid is much less than the capacitance between the plate and cathode. This reduced output capacitance enables a larger inductance to be used to resonate to the desired frequency, which in turn simplifies the construction of the tuned circuits, especially at the higher end of the television band.

Referring to the schematic in Fig. 4-6, the antenna or transmission line is connected across the choke, L_K. This coil is changed for the different television channels in order to present a 300-ohm load to the line for a proper match. In the plate circuit of the r-f amplifier, the autotransformer L_PL_S and the shunting capacitances form a double-tuned coupling circuit. The primary L_P is tuned with the shunting capacitance formed by the output capacitance of the 6AU6 plus the stray capacitance due to wiring. The secondary L_S is tuned with the shunting capacitance consisting of the input capacitance of the 7F8 mixer plus the strays. The circuit is tuned by adjusting the inductance of L_P and L_S and by varying the coupling between them. The windings are overcoupled to provide the necessary bandwidth. For proper alignment, the inductance of either L_P or L_S can be varied by moving a few turns of the coil. The coupling between them can be varied by mov-

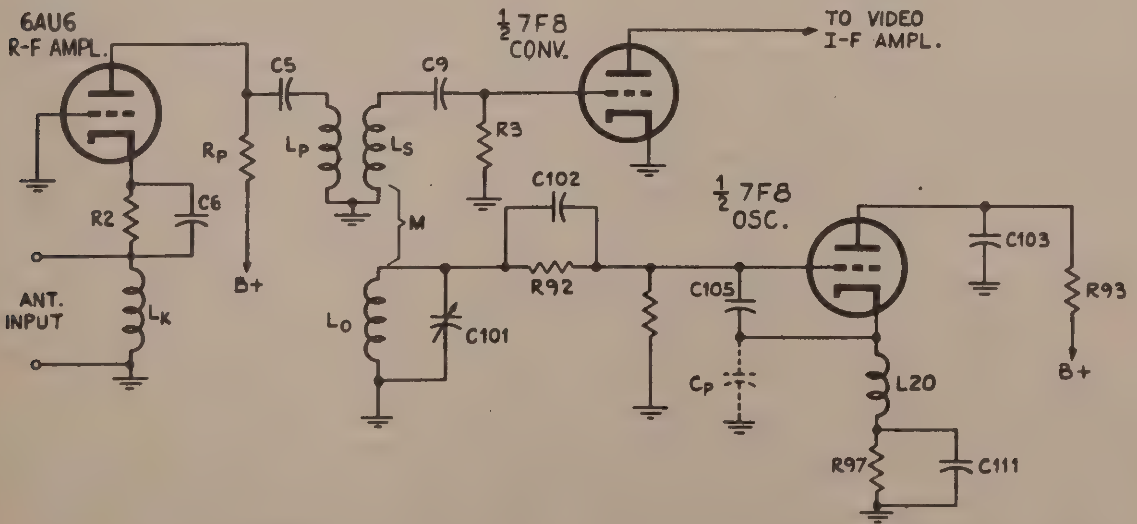


Fig. 4-6.—The r-f amplifier, converter, and oscillator of the GE model 802.

Courtesy General Electric Co.

ing the coils closer together or farther apart on the assembly. On channels 1 and 2, the coupling transformer is triple-tuned to provide greater selectivity. This is shown in Fig. 4-7. The three windings of L_P, L_S, and L_T are all inductively coupled to each other. The third winding L_T is tuned with the shunting capacitance C_T to the same frequency as the primary and secondary windings. The reason greater selectivity is necessary on channels 1 and 2 is that there is possibility of image interference from the f-m band which is from 88 to 108 megacycles. When the oscillator is set for channel 1, for example, the oscillator frequency will be 26.4 mc (the i.f. for this receiver) above the picture carrier of 45.25 mc or 71.65 mc. If a station in the f-m band at 98.05 mc gets through the r-f amplifier, it will beat with the oscillator and also form an i.f. of 26.4 mc. This will go through the i-f amplifiers and cause interference with the picture. This type of interference can be eliminated by improving the selectivity of the tuned circuits before the mixer, and in this receiver it is done by using triple-tuning instead of double-tuning on the television channels 1 and 2. For the other channels, the coupling circuit is as shown in Fig. 4-6, except that loading resistors are placed across the secondary L_S for channels 3 to 6 in order to provide the proper bandwidth.

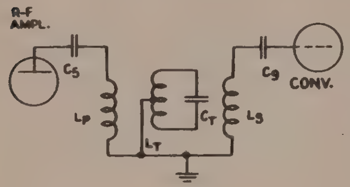


Fig. 4-7. — The triple-tuned coupling circuit used for channels 1 and 2 in the GE model 802.

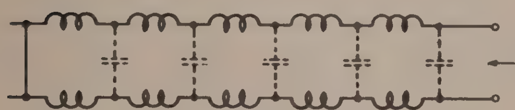
From the coupling circuit the signal is fed to the grid of the mixer tube, which is one half of a 7F8, a dual-triode tube. The other half of the 7F8 is used as the oscillator. The output of the oscillator is also coupled to the grid of the mixer tube because the oscillator coil L_O is wound on the same form as the coils L_P and L_S and is, therefore, inductively coupled to the other windings. The output of the mixer is the difference between the two signals, the i.f. The oscillator in this receiver is a modified Colpitts. The feedback circuit consists of capacitor *C*105 in shunt with the interelectrode capacitance between the grid and cathode of the 7F8 and the interelectrode capacitance between the plate and cathode. Since the plate of the oscillator is at ground for r-f frequencies because of the by-pass capacitor *C*103, the plate-to-cathode capacitance appears between the cathode and ground and is indicated by the capacitance C_P in Fig. 4-6. The oscillator is tuned to the correct frequency by resonance between the coil L_O and the capacitor *C*101. The inductance of the coil L_O can be changed by spreading or compressing the turns of the coil. A different coil is switched into the circuit for each television channel. Since the plate of the oscillator is grounded by capacitor *C*103, the cathode has to be removed from ground. The choke *L*20 places the cathode above ground for r.f. but provides a path to ground for d.c.

RCA 621TS

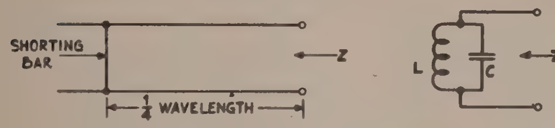
The r-f amplifier, mixer, and oscillator circuits of the RCA model 621TS are assembled on a separate subchassis of the receiver and this unit is also used in the RCA model 630TS. Three 6J6 tubes are used for the r-f amplifier, mixer and oscillator. This tube is a twin-triode miniature tube that has been designed for operation at high frequencies. These tubes are connected in a push-pull arrangement in all input and output circuits except for the output circuit of the mixer, where the plates are connected in parallel to provide a single-ended output. Up to this point, all the plate and grid circuits are operated push-pull. This arrangement is necessary because of the resonant line tank circuits employed. The series of inductances in the plate circuit of the r-f amplifier and oscillator, and in the grid circuit of the mixer make up an artificial transmission line. These small elements of inductance, resistance, and shunting capacitance take the place of the distributed inductance, resistance, and capacitance of an actual transmission line. They are properly proportioned so that the tube sees the same impedance as though it were working into an actual transmission line. Of course, an actual transmission line of the proper length could be used in the same circuit, but the length would make its construction impractical. At the lowest picture carrier frequency, 45.25 mc, a quarter wave-length would be about 4.5 feet long. An artificial line can be made much shorter than this. In the RCA model 621TS, the resonant line is small enough so that it can be constructed around the channel selector switch.

An artificial line made up of series inductances and shunt capacitances is shown in Fig. 4-8 (A). The capacitances shown do not have to be actual capacitors. They can be the capacitances between pairs of inductances. At the right end terminals of this line a certain impedance Z will exist. If the artificial line is properly designed, this impedance will be the same as though the line were an actual transmission line as shown at (B). Therefore, as far as the circuit connected to the terminals is concerned, a line exists. If a shorting bar were placed across the line shown in Fig. 4-8 (B) a quarter wave-length from the terminals, we would have a resonant line which would have the characteristics of a tuned circuit as shown at (C). The resonant frequency of this tuned circuit would be the frequency at which the line is a quarter wave-length long. If the shorting bar can be moved along the line, we can make the line resonant to different frequencies. The resonant frequency would become higher as the line was made shorter and lower as the line was made longer. Thus we can simulate the effect of changing *L* and *C* in the tuned circuit at (C) by merely changing the position of the shorting bar on the line shown at (B). In the artificial line used in the front end of this receiver, the same effect is achieved by switching the shorting bar to different points along the line.

Since a resonant line looks like a tuned circuit at the terminals, the question may be raised as to why a line is used at all. Why not instead use a tuned circuit consisting of a coil and capacitor? One advantage of the line over an ordinary tuned circuit is that the line has a higher impedance at resonance, making possible a higher gain in the amplifier. When used as an oscillator tank, the higher Q of the resonant line will result in increased stability of the oscillator. Since the effect of changing the position of the shorting bar on the line is the same as changing L and C in an ordinary tuned circuit, it is possible to keep the L/C ratio fairly constant and thus have a constant bandwidth over the range of frequencies in the band. The switching arrangement when using a resonant line is simpler, since components do not have to be removed from the circuit to be replaced by other



(A) ARTIFICIAL LINE


SHORTING BAR
Z
¼ WAVELENGTH
L
C
Z

(B) ACTUAL LINE (C) EQUIVALENT CIRCUIT

Fig. 4-8.—The artificial resonant lines used as tank circuits in the RCA model 621TS are shown in (A). The physical line which (A) replaces, is shown in (B) and the equivalent circuit in (C).

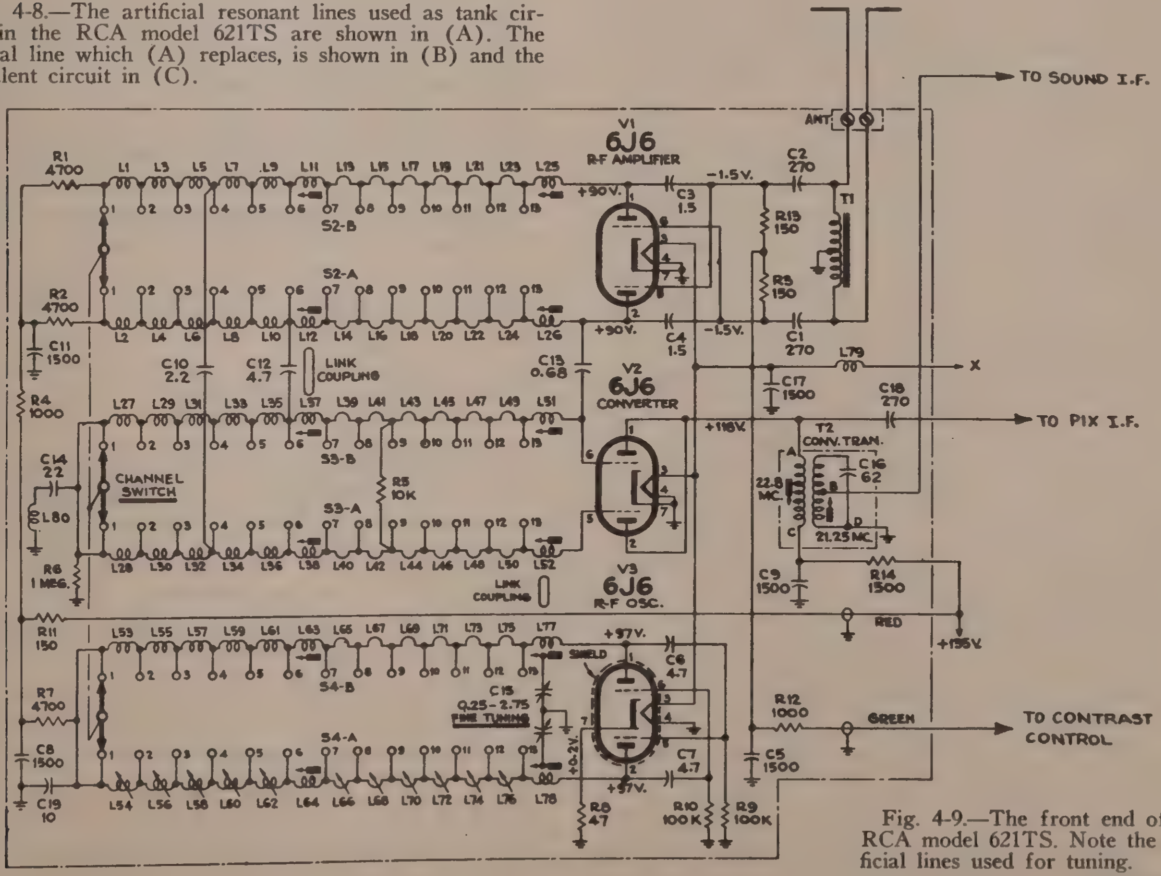

TO SOUND I.F.
ANT.
V1
6J6
R-F AMPLIFIER
R1 4700
C2 270
-1.5V.
+90V.
C3 1.5
T1
R13 150
R3 150
S2-B
S2-A
R2 4700
C1 270
+90V.
C4 1.5
-1.5V.
L79
X
C11 1500
C10 2.2
C12 4.7
LINK COUPLING
C13 0.68
V2
6J6
CONVERTER
C17 1500
C18 270
TO PIX I.F.
R4 1000
+118V.
T2 CONV. TRAN.
C16 62
22.8 MC.
21.25 MC.
C14 22
L80
CHANNEL SWITCH
R5 10K
S3-B
S3-A
R6 1 MEG.
LINK COUPLING
V3
6J6
R-F OSC.
C9 1500
R14 1500
RED
+135V.
R11 150
+97V.
SHIELD
C6 4.7
S4-B
C15 0.25-2.75 FINE TUNING
R7 4700
S4-A
+10.2V.
C7 4.7
R12 1000
GREEN
TO CONTRAST CONTROL
C8 1500
C5 1500
+97V.
C19 10
R8 47
R10 100K
R9 100K

Fig. 4-9.—The front end of the RCA model 621TS. Note the artificial lines used for tuning.
Courtesy RCA

components. All the switch has to do is to place a short at different places along the line to tune to the different television channels.

Referring to the schematic of the front end shown in Fig. 4-9, the transmission line from the antenna, which should be a 300-ohm line, is connected to the grids of the 6J6 r-f amplifier. The resistors *R*13 and *R*3 provide a proper match for the 300-ohm transmission line. The coil *T*1 acts as a bypass for frequencies that are below the television band and effectively shorts them to ground to prevent them from causing interference in the receiver. Since a triode amplifier is used, neutralization is necessary to cancel the effect of feedback due to the interelectrode capacitance between grid and plate which may cause oscillation or instability. This is accomplished by the capacitor *C*3 and *C*4. In the resonant line in the plate circuit, *L*25 and *L*26 are adjustable to provide the correct length of line for the 13th channel. The following inductances *L*13 to *L*23 on the upper side of the line and *L*14 to *L*26 on the lower side are fixed. Physically they consist of a strap between the switch contacts. The shorting bar is moved along this strap to tune to channels 12 to 7. The next coils in the line are *L*11 and *L*12 which are adjustable to take care of the large change in frequency between channel 7 and channel 6. The following inductances, *L*1 to *L*9 and *L*2 to *L*10, are fixed and are switched into the line for tuning to channels 5 to 1.

The resonant line in the mixer grid circuit is similar to the line in the r-f amplifier plate circuit and operates in the same manner. This line is coupled to the r-f line by the coupling capacitors *C*10, *C*12, *C*13 and also by a link. The series combination of *L*80 and *C*14 is tuned to the i.f. to short out any interference from external signals at the i.f. and to prevent feedback from the output of the mixer at the i.f. The grid circuit of the mixer is also coupled to the oscillator by means of a link between the mixer line and the oscillator line. The oscillator line is constructed in the same manner as the r-f and mixer line except that provision is made for tuning the line for each television channel. This is done by varying the inductances *L*54 to *L*78 in the lower leg of the mixer line. *L*63, *L*64, *L*77, and *L*78 are varied by changing the position of an iron core within the coil, as are the variable inductances in the r-f and mixer line. The other variable inductances in the oscillator line are adjusted by changing the position of a brass screw that is placed close to the strap sections, *L*66 to *L*76, or by moving a brass core within the coils *L*54 to *L*62. The effect of the brass core is the opposite of an iron core. The inductance of the coil is decreased when the core is moved farther inside the coil, and increased when the core is moved out of the coil. The feedback circuit for the oscillator is from the plate of one triode to the grid of the other triode by means of the capacitors *C*6 and *C*7. The small variable capacitor *C*15 is the fine-tuning adjustment. This capacitor can shift the frequency above or below the center frequency by a small amount. This range varies from about 300 kc at the lowest television channel to about 750 kc at the highest channel. The control for *C*15 is on the front panel and is mounted on a concentric shaft with the station selector switch.

The signal from the local oscillator and the incoming signal are combined in the mixer tube and a beat frequency is produced which is equal to the difference of these two frequencies. This is the i.f. and appears in the output circuit of the mixer. The mixer transformer *T*2 is not only the plate load of the mixer but also serves to separate the picture i.f. from the sound i.f. The primary of this transformer is tuned with the output capacitance of the 6J6 mixer to 22.8 mc. This frequency is used rather than the center frequency of the picture i-f response because the i-f amplifiers in this receiver are stagger-tuned. The secondary winding on the mixer transformer together with capacitor *C*16 is tuned to the sound i.f. of 21.25 mc. This circuit acts as a trap and removes the sound i.f. from the primary winding by absorbing that frequency. The sound i-f signal that appears on the secondary is coupled to the sound i-f amplifiers, and the picture i.f. on the primary, from which the sound i.f. has been removed, is coupled to the picture i-f amplifiers through the coupling capacitor *C*18.

Du Mont Model RA-103

The r-f amplifier, oscillator, and mixer circuits for the Du Mont model RA-103 are shown in Fig. 4-10. The r-f amplifier uses a 6J6 twin-triode tube with the two triode sections connected in parallel. The input circuit is designed to match a 75-ohm coaxial transmission line. The r-f amplifier employs a grounded-grid type of circuit arrangement, therefore the input signal from the antenna or transmission line is fed to the cathode circuit of the 6J6. The advantages of a grounded-grid amplifier have been discussed previously in this chapter in conjunction with the GE model 802. Referring to Fig. 4-10, the coil *L*106 across the input circuit provides a bypass to ground for any interference below the television band. The grids of the r-f amplifier are returned to ground through the parallel combination of *C*116 and *R*111 rather than directly in order to suppress parasitic oscillations.

The 6J6 r-f amplifier is coupled to the grid of the 6AK5 mixer by means of a double-tuned coupling network. The coils *L*101, *L*102*A* and *L*104-*L*102*B* are tuned with the shunting capacitances due to the output capacitance of the 6J6, the input capacitance of the 6AK5 and the coupling capacitors *C*105, *C*106, and *C*107. The inductance of coils *L*102*A*, *L*102*B* and oscillator coil *L*102*C* is continuously variable by means of a sliding contact on each coil. These are part of a unit known as the Inductuner. The inductance of these coils can be varied sufficiently to cover the entire television band from 44 to 216 megacycles. The inductance variation of each coil is approximately from 0.02 to 1.0 microhenry, which is spread over 10 turns of the Inductuner. This system of tuning has several advantages. The entire television band is covered by one control without the use of separate tuned circuits for each channel. The bandwidth is fairly constant over the entire range of frequencies. The f-m band of 88-108 mc is covered in the continuous tuning range of the Inductuner which simplifies the construction of combination television and f-m receivers.

The resistors *R*110 and *R*104 load down the tuned circuits in order to provide sufficient bandwidth, which should be about 6 mc in order to pass both the picture and sound signal. The capacitors *C*105 and *C*106 can be adjusted for proper alignment. The output of the local oscillator is also coupled to the grid of the 6AK5 mixer tube by means of the coupling capacitor *C*112. The two signals beat together in the mixer and produce the i.f. in the output of the 6AK5. This frequency is equal to the difference in frequency of the local oscillator and the incoming signal to which the receiver is tuned. In this receiver the picture i.f. is 26.4 mc and the sound i.f. is 21.9 mc. Both these frequencies are fed to the input of the first i-f amplifier, to be removed farther on in the circuit.

The oscillator circuit is a modified Colpitts and uses one half of a twin-triode 6J6. The feedback consists of the interelectrode capacitance between plate and cathode and between grid and cathode of the 6J6. The frequency of the oscillator is determined by the resonant frequency of the parallel LC tank in the grid-plate circuit. The inductance of *L*102*C* is continuously variable and is mechanically coupled to the other coils in the Inductuner. The trimmer capacitor *C*111 and the inductance *L*103 provide proper tracking of the oscillator. Bias on the oscillator tube is provided by the combination of *C*114 and *R*109. The plate and grid of the second section of the 6J6 are grounded since this section of the tube is not used. The capacitor *C*115 across the heater of the oscillator places the heater at ground for r.f. Notice that the r-f amplifier and converter heaters are also bypassed with a similar capacitor. This prevents coupling between these three stages through the common heater supply and also prevents interfering signals from the rest of receiver. The plate voltage supplies are also separately bypassed at each tube for the same reason.

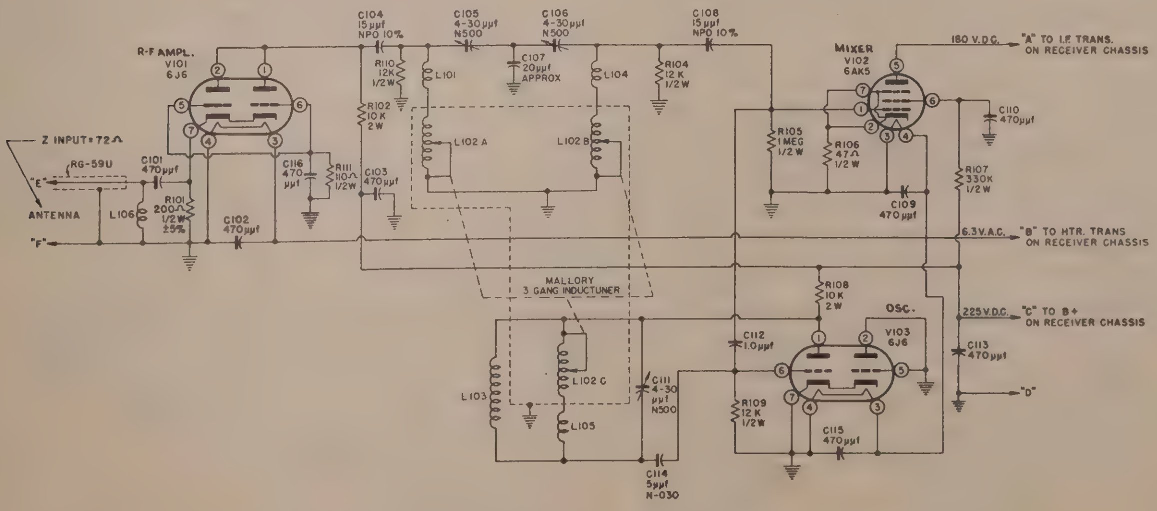

R-F AMPL.
V101
6J6
Z INPUT=72Ω
RG-59U
"E"
ANTENNA
"F"
C101
470μμf
L106
R101
200Ω
1/2W
±5%
C102
470μμf
C116
470
μμf
R111
110Ω
1/2W
C103
470μμf
C104
15μμf
NPO 10%
R110
12K
1/2W
R102
10K
2W
L101
C105
4-30μμf
N500
C106
4-30μμf
N500
C107
20μμf
APPROX
L104
L102 A
L102 B
C108
15μμf
NPO 10%
R104
12 K
1/2 W
MIXER
V102
6AK5
R105
1MEG
1/2 W
R106
47Ω
1/2 W
C109
470μμf
C110
470μμf
R107
330K
1/2 W
180 V.D.C.
"A" TO I.F. TRANS.
ON RECEIVER CHASSIS
6.3V.A.C.
"B" TO HTR. TRANS
ON RECEIVER CHASSIS
225V.D.C.
"C" TO B+
ON RECEIVER CHASSIS
"D"
C113
470μμf
MALLORY
3 GANG INDUCTUNER
L103
L102 C
L105
C111
4-30
μμf
N500
C114
5μμf
N-030
C112
1.0μμf
R108
10K
2W
OSC.
V103
6J6
R109
12 K
1/2W
C115
470μμf
1
2
3
4
5
6
7

Fig. 4-10.—The r-f amplifier, oscillator, and mixer circuits of the Du Mont model RA-103.

Courtesy of Allen B. Du Mont Laboratories, Inc.

Westinghouse Model H-181

The local oscillator circuit used in the Westinghouse model H-181 is shown in Fig. 4-11. The frequency stability of this oscillator is improved by the use of an automatic frequency control (afc) circuit. In Fig. 4-11, the first half of the 6J6 tube is used as the local oscillator, and the second half of the 6J6 is used as a reactance tube in the afc circuit. The oscillator circuit is a modified Colpitts that employs the interelectrode capacitances of the tube as the feedback circuit. The cathode of the oscillator tube V_1 is grounded for r.f. by capacitor C_4, thereby effectively placing the cathode at the junction of the plate-to-cathode and grid-to-cathode capacitances of the tube. The frequency of oscillation is determined by the resonant frequency of the tank circuit consisting of the oscillator coil L_1, capacitors C_2 and C_3, the interelectrode capacitances of V_1, and the shunting capacitance C_s of the reactance tube V_2. The oscillator is set for the correct frequency for each television channel by switching in different values of inductance for L_1. The frequency of the oscillator can also be changed by changing the value of any capacitance in the tank circuit. If this could be done automatically as the oscillator drifts, due to changes in voltage or temperature, and if the change in capacitance were sufficient and in the right direction, the frequency of the oscillator could be kept constant. This is the function of the reactance tube V_2. The shunting capacitance C_s supplied by this tube varies in such a manner as to keep the frequency of the oscillator constant.

To analyze the operation of the reactance tube, let us first examine the circuit shown in Fig. 4-11. The plate of the reactance tube is connected to the plate of the oscillator tube. The choke L_2 keeps the two plates from being grounded for r.f. through the B+ supply. The two cathodes are tied together and cathode bias is obtained by the voltage divider consisting of resistors R_1 and R_2. This bias holds the reactance tube at the correct operating point. The oscillator uses grid-leak bias supplied by the combination of C_3 and R_3 and is not affected by the positive voltage on the cathode because the grid of the oscillator is returned to the cathode rather than ground by the resistor R_3. The grid of the reactance tube is coupled to the oscillator coil through the network consisting of R_4, C_5, C_6, and the input capacitance C_i of tube V_2. The cold grid-to-cathode capacitance of the 6J6 is only about 2 $\mu\mu f$, but due to the gain of the tube, the input capacitance of the reactance tube will appear to be much larger than this (Miller effect). The capacitor C_5 will have very little reactance at the operating frequencies of the oscillator (approximately from 70 to 240 megacycles) and resistor R_4 may be considered to be in parallel with capacitor C_6. This capacitor enables the reactance tube to operate properly over this large range of frequencies.

The current flowing through C_i will be determined mainly by resistor R_4. This current will therefore be in phase with the voltage producing it, that voltage on the grid of V_1. The voltage across C_i will lag the current by 90° and, therefore, lag the grid voltage of V_1 by 90°. The plate current of V_2, which is in phase with the grid voltage on V_2, will also lag the voltage on the grid of V_1 by 90°. Since V_1 is an oscillator, the voltage on the plate will be 180° out of phase with the voltage on the grid, and since the two plates are tied together, the plate voltage of V_2 will also be 180° out of phase with the grid voltage of V_1. We have already shown that the plate current of V_2 lags the grid voltage of V_1 by 90°. Therefore, the plate current of V_2

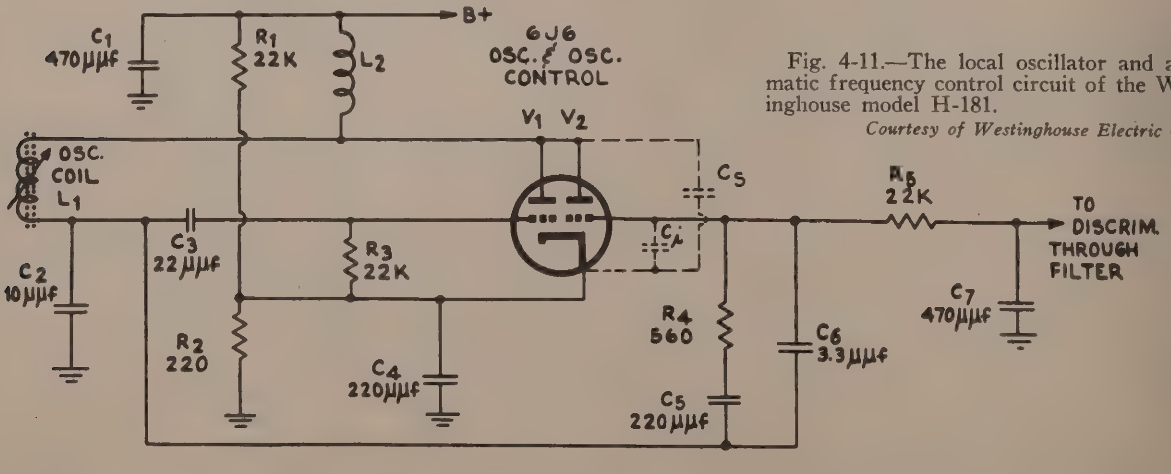


Fig. 4-11.—The local oscillator and automatic frequency control circuit of the Westinghouse model H-181.

Courtesy of Westinghouse Electric Corp.

will lead the plate voltage of V_2 by 90°. This is the same relationship that would exist with a capacitance in the circuit, and because of this, tube V_2 will appear as a capacitive reactance, or an equivalent capacitance C_s, across the plate circuit of the oscillator V_1. The value of this capacitance can be varied in the following manner. If the grid of V_2 is made more positive, the plate current of this tube will be increased. This would make the tube appear as a smaller reactance, and since the reactance is capacitive, the equivalent capacitance C_s appears larger. Conversely, if the grid were made more negative, the capacitance C_s would be decreased.

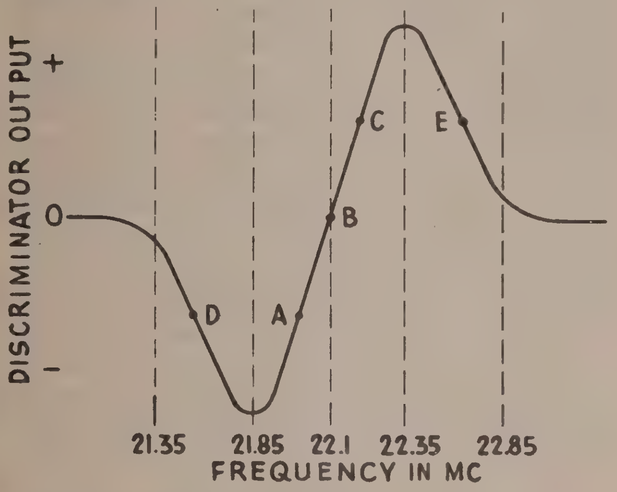


Fig. 4-12.—The discriminator characteristic of the Westinghouse model H-181.

The grid of reactance tube V_2 is connected to the output of the discriminator through a filter which removes the audio component but allows the d-c voltage to be applied to the grid. The discriminator characteristic of this receiver is shown in Fig. 4-12. At 22.1 mc, which is the sound i.f., the output of the discriminator is zero. For frequencies below this value, the output will be negative, and for frequencies above, the output will be positive.

Let us consider three possible conditions that may arise in the operation of this system. Assuming that the receiver is tuned to channel one, the local oscillator will be operating at a frequency of 71.85 mc. This will beat with the audio carrier of 49.75 mc to produce a sound i.f. of 22.1 mc. This will correspond to point *B* on the response characteristic in Fig. 4-12. The output of the discriminator will be zero, and no correction will be applied to the oscillator. The second case is that where the oscillator has drifted to a higher frequency due to some change in voltage or temperature. This results in an audio i.f. which is too high and corresponds to point *C*. The output of the discriminator is then positive, and this positive voltage applied to the grid of the reactance tube makes the capacitance C_s larger, as explained previously. This larger capacitance makes tube V_1 oscillate at a lower frequency, bringing the audio i.f. back to the proper frequency at point *B*. In the third case, the oscillator drifts to a frequency below 71.85 mc. This would result in a sound i.f. which is below 22.1 mc, as shown by point *A* on the discriminator characteristic. The negative voltage produced by the discriminator and applied to the grid of the reactance tube would decrease the capacitance C_s. This would make the oscillator work at a higher frequency and bring the sound i.f. up to point *B*. From the foregoing, it can be seen that the afc system will tend to keep the oscillator at the correct frequency for the particular channel being received at all times, despite any tendency of the oscillator to drift due to changes in voltages or temperature. However, the range of control of this system is limited to the frequency range between the peaks of the discriminator characteristic. This is about 500 kc in this receiver. If, for any reason, the oscillator should drift past the peak on either side, for example, point *D* or *E*, the afc system would lose control and not restore the oscillator to the proper operating frequency. This happens because the control voltage at these points decreases with increasing drift in frequency, instead of increasing as it does at points *A* and *C*.

To get the maximum advantage from the afc system, it is important that the initial operating condition of the oscillator be at point *B*. This adjustment is made in the alignment of the receiver by disabling the afc system and adjusting the frequency of the oscillator until the output of the discriminator is zero. This adjustment should not be made until the receiver has reached normal operating temperature, which usually takes at least 20 minutes.

CHAPTER 5

THE F-M SOUND CHANNEL

By Seymour D. Uslan

In the first type of commercial television receivers the sound signals were amplitude modulated and the sound channel contained circuits similar to the a-m superheterodyne receiver. Today the sound signals used with television are frequency modulated. There are many reasons why f-m sound broadcasting is superior to amplitude modulation in this field. Reduction in interference, better high fidelity sound, possibly better signal-to-noise ratio are just a few of the advantages of f.m.

Due to the nature of f-m and video signals, they can both be received by the same antenna. The circuits of the f-m sound section of a television receiver are very similar to those of the regular f-m receivers that now appear on the market. The circuits that we are essentially interested in are the i-f section, the detector section, and the audio section. These three topics of the f-m sound channel will now be considered in detail.

THE I-F SYSTEM

The output signal from the converter or mixer from the front end of the set is a combined video and sound i-f signal. In some television receivers the sound i.f. is taken from the output of the mixer or converter tube in the front end of the set. In other receivers the combined i-f signal output from the converter or mixer is impressed across an amplifier. This was already discussed in connection with the front end of the set. This i-f amplifier is usually called the first video or pix i-f amplifier. In this case the sound i.f. is taken from the output of this tube through a sound i-f tuned circuit.

In either type of receiver the sound i-f signal is impressed across the grid of the first sound i-f amplifier. From this point on, the analysis of the f-m sound section of the television receiver is very similar to a regular f-m receiver. The only major difference between them lies in the frequencies involved.

Frequency Deviation

In regular f-m broadcasting the maximum bandwidth involved, as defined by the FCC, is equal to 100 kc on either side of the center frequency. This is due to the 75-kc peak frequency deviation for 100-percent modulation plus the 25-kc guard band. In television the sound transmitter according to FCC standards should be designed to operate satisfactorily with a peak frequency deviation (frequency swing) of 25 kc for 100-percent modulation. In other words we have a 150-kc peak-to-peak frequency deviation in regular f-m broadcasting and a 50-kc peak-to-peak frequency deviation in television sound transmission, both for 100-percent modulation. Consequently we find that the f-m sound signal in television has 1/3 of the frequency deviation of a regular f-m broadcasting transmitter for 100-percent modulation.

This reduced deviation and, hence, bandwidth does not put any limitation on the range of audio frequencies that can be passed. As in regular f-m broadcasting, the f-m sound transmitter in television has to be designed to pass audio frequencies from 50 to 15,000 cycles. Therefore high fidelity in the television sound section is still possible even though the peak frequency deviation of output signal is at a maximum of 25 kc. The primary difference between a wide bandwidth and a smaller bandwidth is that in the latter the signal-to-noise ratio will be somewhat lower, considering the same amount of amplification in the receiving systems. This can be remedied, if necessary, by increasing the strength of the signal in the i-f stages of the sound channel of the television receiver.

The FCC recommends that the television sound transmitter be designed to operate satisfactorily with a peak frequency deviation of at least 40 kc. This allows for slight amounts of overmodulation in the transmitter.

The I-F's and Image Frequency Interference

The intermediate frequencies of the sound channel of some of the first television receivers employ-

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Note: For a more detailed explanation of frequency modulation, see "FM Transmission and Reception" by Rider and Uslan.

ing f-m sound was in the vicinity of 8 mc. Today many of these frequencies are much higher—somewhere around 21 and 22 mc. Some of the common sound i.f.'s used today are 21.25 mc, 21.9 mc, and 22.25 mc. The i.f. used in most of the f-m receivers on the market is 10.7 mc. This intermediate frequency, in accordance with the commercial f-m band of 88 to 108 mc (20-mc bandwidth), does not allow image frequency interference. In other words, if twice the i.f. employed is greater than the bandwidth of the frequency range in question, image frequency interference cannot occur.

The bandwidth of frequencies involved in the 13 television channels is somewhat different. In this case we can consider these channels as divided into two groups—those channels from 1 to 6 in one group and those those from 7 to 13 in the other. The television sound carrier frequency of the first channel (44 to 50 mc) is equal to 49.75 mc and that of the sixth channel (82-88 mc) is equal to 87.75 mc. This means that the frequency range of the sound carrier signals in the first group, between the first and sixth channels, is 87.75 less 49.75 mc or 38 mc. The sound carrier frequency of the seventh channel (174-180 mc) is 179.75 mc and that of the 13th channel (210-216 mc) is equal to 215.75 mc. Thus the frequency range of sound carrier signals in the second group, between the seventh and 13th channels, is 215.75 less 179.75 mc or 36 mc.

From this analysis we find that the frequency range of the sound carrier signals is 38 mc in the first group and 36 mc in the second group. Consequently, if twice the sound i.f. is greater than the largest of these frequency ranges, then image frequency interference will not result. Thus if the sound i.f. is around 20 mc (as it is today), image frequency interference occurring in either channel grouping is not possible. The 86-mc separation between the sixth and seventh channel does away with the possibility of image frequency interference between channels.

This same analysis also holds true for the video i-f system regarding the picture carriers causing image frequency interference within the video band.

If harmonics of the sound i.f. are strong enough, they may cause interference in the video channel if picked up by the r-f section of the receiver. For instance, if a television receiver with a sound i.f. of 21.25 mc is tuned to channel 3 whose frequency range is 60 to 66 mc, then the third harmonic of this i.f., which is 63.75 mc, can be picked up by the r-f section and passed on through the video system. If this happens, the third harmonic of the i.f. will be detected by the video detector and passed along with the video signal to the picture tube, interfering with the pattern on the picture tube.

The fourth harmonic of this same sound i.f. can cause interference on channel 6 whose frequency range is 82 to 88 mc. This is so because the fourth harmonic of 21.25 mc is 85 mc which falls in the center of the 6th television channel. In similar fashion if the sound i.f. is 22.25 mc, its second harmonic can cause interference on the second television channel and its third harmonic on the fourth channel.

I-F Stages

The primary function of the i-f amplifiers and transformers in the sound section of a television receiver is the same as in any ordinary receiver. The i-f amplifiers provide a large part of the necessary gain and selectivity needed for the proper operation of the sound system. Even though the peak-to-peak frequency deviation for 100-percent modulation in television sound broadcasting is 50 kc, the bandpass characteristics of many of the i-f transformers are greater than the 50 kc. One of the reasons advanced for this large bandwidth is so the transformer will accept the i.f. even when the r-f oscillator of the set drifts in frequency also causing the i.f. to drift. In other words, this 50-kc bandpass is the miminum bandwidth requirement for the i-f transformer.

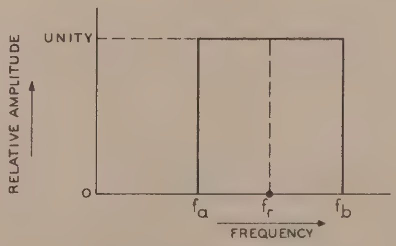


Fig. 5-1.—An ideal selectivity curve that is flat-topped would pass all frequencies between the limits f_a and f_b with equal amplitude; such a curve cannot be realized in practice with the use of a single i-f transformer network.

The nature of a true f-m signal is such that it varies only in frequency but is constant in amplitude. The f-m signal input to the i-f stages for proper operating conditions should be constant in amplitude and it is desired that the output also be constant in amplitude. For the output f-m signal from the i-f system to be constant in amplitude, the response of these i-f stages should be as broad and flat as possible. A broad selectivity curve with a somewhat level or flat characteristic can be achieved but the curve is not perfectly flat-topped and will cause amplitude variations in the f-m output signal from the i-f stages.

A response curve is, after all, a curve that shows the selectivity characteristics of a particular resonant circuit, but these curves are not restricted to operate on a-m or f-m signals. A particular resonant circuit, like a simple parallel LC circuit, will respond to both a-m and f-m signals. This is the case with all i-f transformer networks. The mean resonant frequency and not the type of modulation determines the center frequency of the signal that can be received.

The ideal selectivity characteristic for an f-m signal would be a flat-top response curve as illustrated in Fig. 5-1, where only frequencies between f_a and f_b would be accepted, and all would have the same amplitude. However, such a curve cannot be realized in practice with the use of a single i-f transformer network. Methods which help approach the shape of such a curve entail the use of more than one i-f transformer network, and preferably of three.

Manufacturers are cognizant of the fact that such a curve is only ideal and is extremely difficult to obtain. They are faced with the necessity of getting a final output i-f signal that is high enough in gain and broad enough to pass at least 50 kc at the half-power points (also called the 3-db points) of the i-f response without any discrimination. They are also concerned with the fact that a constant amplitude signal is preferable, but since limiters and special detection circuits that effectively do away with the amplitude variations are available, such design is of a secondary nature. A picture of a typical response characteristic of i-f stages as used in the sound section of television receivers is shown in Fig. 5-2. From the shape of this curve we can readily see that the f-m i-f signal, even though constant in amplitude at the input to the stage, will at the output be an f-m i-f signal that is varying in amplitude as well as in frequency. The shape of the response curve determines the shape of the amplitude variations of the output f-m signal.

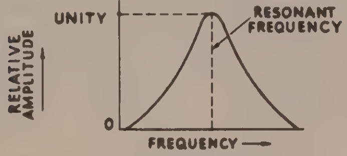


Fig. 5-2.—Typical response characteristic of a sound i-f stage in a television receiver.

In f-m broadcast receivers, the desired selectivity of the i-f stages is much broader than that of the television sound channel. To obtain the necessary selectivity in the i-f section of f-m broadcast receivers, various methods are used to broaden the selectivity. Some methods used are to employ overcoupled i-f transformers or a series of stagger-tuned i-f transformers. This latter method is employed in the video i-f sections of many television receivers and will be discussed in the video section later on.

In the sound i-f section of television receivers, the usual method is to employ two i-f amplifiers in conjunction with three i-f transformers where the transformers are all single peaked to the same resonant frequency. The single peaking is usually somewhat under critical coupling. By the use of two amplifiers and sometimes more, the necessary gain for the particular set is thus obtained. Even though the higher i.f. used in television sound systems causes a lower gain per stage, the type and number of tubes used today make it possible for receivers to withstand this loss and still produce the necessary gain.

Detection of the F-M Signal

The f-m sound channel of a television receiver is similar to a regular f-m or a-m broadcast receiver in that the i-f stages are followed by a system of detection which is followed by an audio system. The primary difference between a-m and f-m receivers is in the methods of detection. The detectors for a-m and f-m differ considerably in circuit construction and performance but their purposes are essentially the same—to detect or demodulate their input modulated signal to obtain the audio modulation component.

The f-m detector system must respond only to frequency variations and not to amplitude variations. This is essential because the signal resulting from detection should contain only those audio characteristics of the modulating signal present at the sound transmitter. Television receivers today employ two main types of f-m detectors in the sound channel—the Foster-Seeley, or phase discriminator, and the ratio detector. The latter type of circuit is newer and only one tube is employed to accomplish the process of f-m detection and amplitude limitation. The former type circuit employs one tube for amplitude limitation and another one for detection. A separate tube is used because the phase discriminator will respond to amplitude variations as well as frequency variations. These amplitude variations, whether small or large, are caused primarily by the response characteristics of the i-f transformers and also by superimposed noise.

These two types of detector systems are so important to the proper operation of the television sound channel that they will be considered in detail here. We will analyze their circuits from a general viewpoint and then show possible modifications of such circuits.

The first type of f-m detector system that was used commercially in f-m receivers was the phase discriminator where a separate amplitude limiter was employed. In this system the limiter tube, which precedes the discriminator detector, levels off practically all of the a-m variations by clipping off the upper and lower portion of the f-m signal so that it will be constant in amplitude when it reaches the discriminator network.

We will discuss the limiter-discriminator system first and then the ratio detector systems.

THE LIMITER SYSTEM

The limiter stage of the sound channel of a television receiver immediately follows the final i-f stage and in circuit arrangement appears very similar to an i-f stage. In fact, this limiter system may be considered as the last i-f stage of the receiver, although the purpose is one of amplitude limitation rather than amplification.

It was stated that for the i-f response curve not to contribute amplitude variations to the f-m signal, it would have to be flat topped with straight sides as in Fig. 5-1. It is, therefore, seen that the f-m signal, although devoid of a-m variations as it leaves the transmitting antenna, will contain amplitude variations at the output of the last i-f stage. It is the purpose of the limiter to eliminate these amplitude variations.

Ideally the function of the limiter is shown in Fig. 5-3 with respect to an f-m signal. The input to the limiter is an f-m signal that is varying in amplitude as well as in frequency. These amplitude variations are undesired and, by the action of the limiter, are "clipped off" and the output is an f-m signal of constant amplitude.

For an amplifier to act as a limiter, the potentials on the tube are so chosen that the tube will overload easily with a small amount of signal input. A number of special operating conditions are combined to make the limiter function properly. First, the amplifier tube used is usually of the sharp cutoff type like the 6SJ7 or 6SH7 tubes or the miniature type 6AU6 tube. Low values of screen and plate voltage and little or no initial control grid bias are applied to the tube, so that it will quickly overload and plate current cutoff will be rapidly reached.

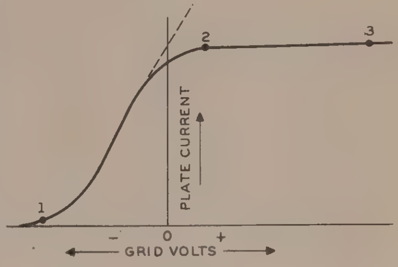


Fig. 5-4. — The limiter plate current–grid voltage characteristic differs from the usual characteristic in that it flattens out in the positive region (beyond point 2). Amplification will occur between 1 and 2; beyond 2, limiting action will occur.

To understand the operation of the limiter fully, let us first examine a typical plate current-grid voltage (i_b-e_c) curve, as illustrated in Fig. 5-4. In general, i_b-e_c curves show the amount of plate current that will flow for a given value of grid voltage. Most i_b-e_c curves take on the shape shown between points 1 and 2 in Fig. 5-4., but between points 2 and 3 this curve differs from most curves typifying amplifier action. A great deal of this difference is brought about by "clipping" parts of the positive halves of the input signal due to the grid and cathode of the limiter acting as a diode at this part

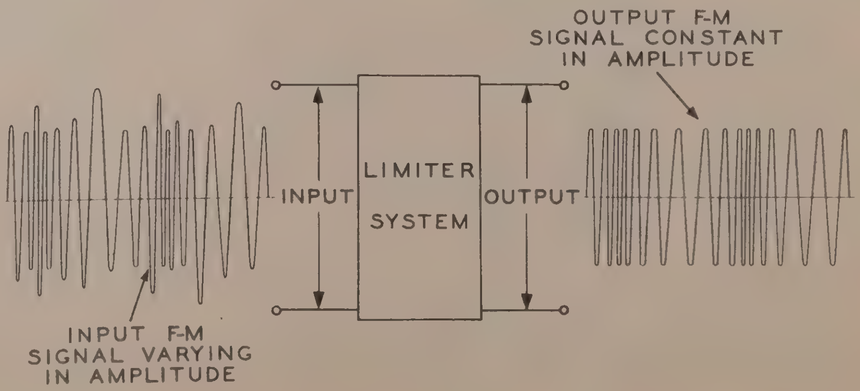

OUTPUT F-M SIGNAL CONSTANT IN AMPLITUDE
INPUT
LIMITER SYSTEM
OUTPUT
INPUT F-M SIGNAL VARYING IN AMPLITUDE

Fig. 5-3.—The action of the limiter in the sound system of a television receiver is to clip off the amplitude variations of the input signal, making its output constant in amplitude.

of the input signal. This is commonly known as diode clipping and will be discussed in detail later on. If diode clipping did not occur, the curve instead of taking on the shape between points 2 and 3, would continue onward as shown by the dashed line.

In the region of the curve between points 1 and 2, the tube, by virtue of the shape of this curve, will act as an amplifier to any input voltage that has values lying in between these two points. Beyond point 2, the curve is seen to level off and, no matter how high the input grid voltage, the plate current will be virtually constant in value. Beyond this point on the curve the tube will function as a limiter. To make sure that the tube functions as a limiter in ideally producing a constant output signal, the instantaneous input grid signal must always rise at least to the value at point 2. If an input signal is such that it does not rise beyond point 2, any amplitude variations in the input signal will be retained in the output. Similarly, if the input signal swings below point 1, its negative peaks will be clipped, because below point 1 the plate current is substantially zero.

Further analysis of Fig. 5-4 brings a few pertinent facts to light. To get limiting action at both extremes of the input signal, this signal must have a swing that falls outside points 1 and 2 on the curve. Any input signal that has amplitude variations not exceeding points 1 or 2 in voltage will, as a consequence, be reproduced in the plate circuit with these amplitude variations and this is undesired. This simple analysis thus reveals that a certain threshold of input voltage to the limiter grid is needed for limiting action at both extremes of the input cycle.

Further examination of Fig. 5-4 must lead to the conclusion that such action of the limiter tube results in the development of distortion in the plate circuit. Plate-current variations are enlarged reproductions of grid-voltage variations. The distortion does exist, but is of no consequence because, when we clip the amplitude of the wave, we do not change the relative frequencies present in the frequency deviation. An instantaneous frequency of say 21.25 mc does not undergo any change if it is increased from 10 volts to 50 volts or reduced from 50 volts to 30 volts. Any harmonics introduced by the clipping section are of no importance, because the frequencies representing these harmonics are outside the range of the resonant circuit in the plate circuit of the limiter, a typical circuit of which appears in Fig. 5-5. This circuit responds only to the range of frequencies representing the frequency deviation on both sides of the carrier, and perhaps a little beyond these limits. The harmonic frequencies are filtered out of the circuit by the tuned transformer which couples the limiter to the discriminator.

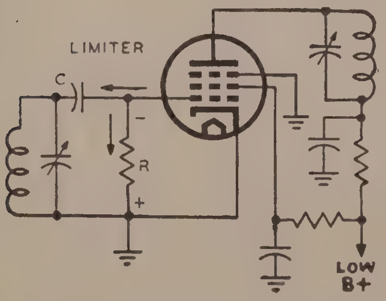


Fig. 5-5—The *RC* combination determines how quickly grid bias on the limiter tube will occur, thus causing automatic bias regulation.

The *RC* network in the grid circuit in conjunction with the low potentials on the tube help provide for proper limiting action of the tube. This will now be discussed in detail.

Analysis of Limiting Action

To understand the true function of a limiter stage we should first know something about its circuit arrangement. Most limiter circuits are very much alike with the chief difference in their grid circuit arrangements. In most grid circuits, however, a grid-leak resistor and capacitor arrangement is used. This *RC* combination makes it possible for the bias on the tube to change in accordance with the diode clipping action of the input signal. The time constant of the *RC* network determines how quickly the grid bias change will occur. (The time constant in seconds is determined by the product of *R* in ohms and *C* in farads.) In the typical limiter circuit illustrated in Fig. 5-5 resistor *R* and capacitor *C* represent the grid bias network. In other receivers the grid resistor *R* is shunted across *C* and both are placed in the grid return circuit. Numerous versions of these two circuits exist, but the analysis to follow with respect to the limiter of Fig. 5-5 will hold for limiter circuits in general.

The grid arrangement, in conjunction with low plate and screen voltage and no fixed bias, causes *automatic bias regulation* of the limiter tube. At any instant of time there is a bias on the grid, but this bias is changing during the first few cycles of input signal. The instantaneous value of the grid voltage determines the operating bias at each instant of time during these first few cycles. After a certain amount of time has elapsed, a point will be reached where the bias on the tube will remain constant with constant input signal voltage. The control grid and cathode of the tube, in conjunction with the *RC* network, produce the bias. The actual plate current-grid voltage curve for this circuit is illustrated in Fig. 5-6. The bias on the tube without any input signal to the control grid is zero volt. In accordance with the potentials on the plate and screen, a certain amount of d-c plate current flows through the circuit. As seen by the curve of Fig. 5-6, plate current cutoff will occur when the grid has a bias of —8 volts.

Since there is zero fixed bias on the tube, the control grid will be driven positive during the positive half of the input signal with respect to the cathode, which is at ground potential. As soon as the grid is driven slightly positive, the grid and cathode act as a diode rectifier where the grid takes the place of the diode plate and grid current starts to flow.

As the grid draws current, the coupling transformer from the preceding stage is effectively loaded so that signal voltage applied directly to the grid can become only slightly positive. This is shown in Fig. 5-7 where the positive parts of the input above the zero-volt line are small due to clipping action. Contrast these true small positive swings shown by the solid lines above the zero grid-voltage axis with the dashed curve. This latter curve shows what the signal voltage would be if there were no clipping due to diode action.

The moment that grid current begins to flow, a charge is stored on the grid capacitor *C*, and, as the signal becomes more positive, the charge on the capacitor increases. On the negative half cycle of input signal, the grid no longer will draw current and the capacitor *C* will begin to discharge through the resistor *R* as shown by the current arrow through *R* in Fig. 5-5. An automatic bias will be developed on the tube, as seen in Fig. 5-7, due to

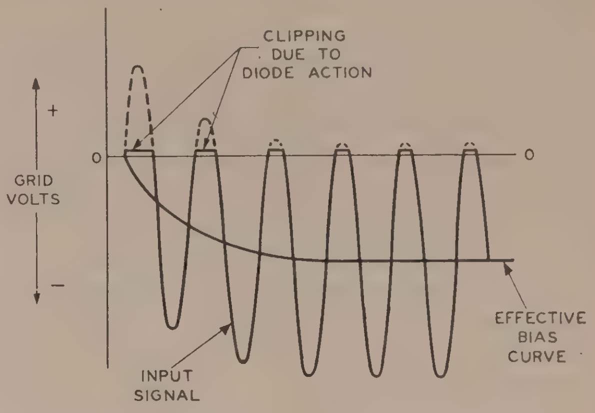


Fig. 5-7.—The solid curves above the zero axis show a uniform amplitude due to clipping; the dotted portions above the zero axis show what the signal would be if no clipping occurred.

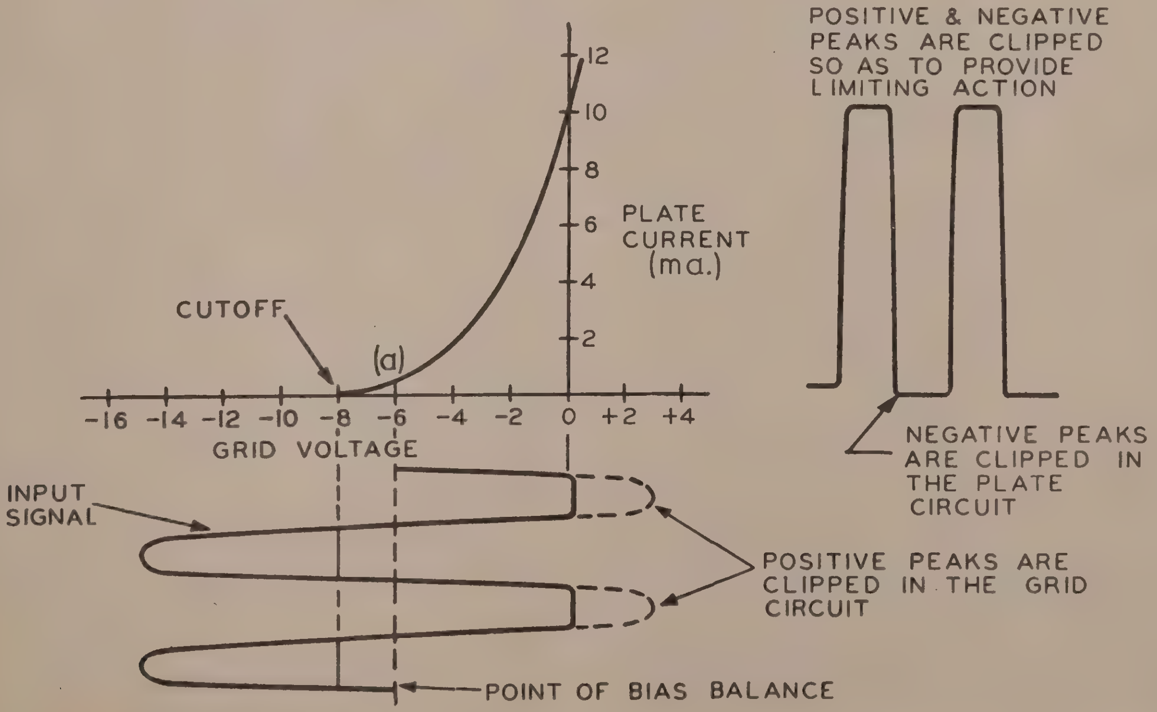


Fig. 5-6.—The clipping action of the input signal to the limiter circuit of Fig. 5-5 is illustrated in the characteristic curve shown here.

circumstances, it will be seen from the vector diagram of Fig. 5-13 that the resultant vector OE_4 representing that voltage across diode D_2 and resultant vector OE_5 representing that across diode D_1 are no longer equal, but that vector OE_4 is greater than vector OE_5. In this instance diode D_2 will draw the greater current, and in Fig. 5-11 load resistor R_2 will have a greater voltage drop than resistor R_1; hence a differential voltage will exist across point *A* to *B*, with point *B* being more positive than point *A*. This is the same as saying point *A* is negative with respect to point *B*.

Applied Frequency Lower Than Resonance

When the instantaneous value of the f-m signal input to the discriminator circuit is such that it is less than the resonant frequency of the discriminator transformer, the differential voltage will still exist across the diode loads, but the polarities will be reversed. Let us see how this happens.

We still are at off-resonance conditions, even though we are on the lower side of the resonant frequency, and the same 180° phase relationship between *E* and E_1 exists. When the applied frequency is lower than that of the resonant frequency of the i-f transformer, the impedance of the secondary of the i-f transformer is such that the capacitive reactance more than balances out the inductive reactance, and the secondary is primarily capacitive. Since this circuit is capacitive, the induced current *I* leads the induced voltage *E*. If the off-resonance conditions are such that a phase angle of 35° again exists between *I* and *E*, *I* will be *leading* *E* by 35°, as seen in the vector diagram of Fig. 5-14. Since the 90° phase relations between voltage E_2 and current *I*, and voltage E_3 and current *I* must still exist, these two voltages are effectively shifted in phase 35° in a counterclockwise or positive direction. This is indicated in Fig. 5-14 where vectors OE_2 and OE_3 are still 180° out of phase with each other, but no longer 90° out of phase with vector OE_1. Now when the individual i-f voltages across the diode circuits are combined vectorially, it will be seen that resultant vector OE_5 applied to diode D_1 is greater in magnitude than resultant vector OE_4 applied to diode D_2. Therefore, the current in the circuit of diode D_1 is greater than the other diode current. This means a greater voltage drop exists across R_1, the load resistor of diode D_1, than across R_2, the load of D_2, and a differential voltage exists between points *A* and *B* of the diode circuit of Fig. 5-11. However, under these conditions the polarity of point *A* will be more positive with respect to the junction of R_1 and R_2 than point *B*. That is, point *B* is negative with respect to point *A*.

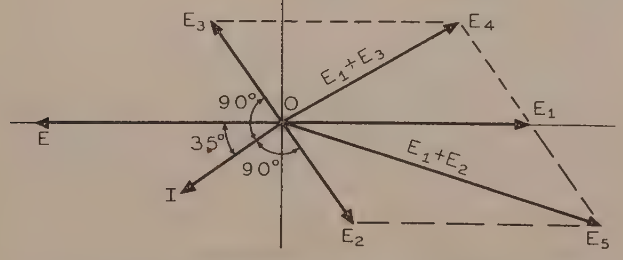


Fig. 5-14.—When the instantaneous frequency of the f-m signal is lower than the resonant frequency of the transformer, the current I will lead the induced voltage E. Compare with Figs. 5-12 and 5-13.

Summary

Summarizing the action described, it is evident that if a varying frequency input signal (one which varies in frequency around a mean) is applied to the discriminator network—provided that the range of frequencies covered is not beyond the acceptance bandwidth of the discriminator transformer—an output signal which changes in amplitude and polarity will be obtained. The frequency swing of the f-m signal is determined by the amplitude of the audio modulating signal. The greater the amplitude, the wider the frequency swing, and the smaller the amplitude, the lesser the frequency swing. The frequency of the audio modulating signal determines the time rate change of the frequency swing. This frequency swing, usually called *frequency deviation*, varies an equal amount on either side of the center or carrier frequency of the f-m signal for each cycle of audio modulating signal. If the audio modulation is a sine wave of constant amplitude or any similar signal of constant amplitude, the frequency deviation will be the same for every cycle of modulation. However, in speech or music the audio modulation is anything but constant in amplitude and, therefore, the frequency deviation for each cycle is usually different.

The output signal from the discriminator is determined by the amount of frequency deviation; the less the frequency deviation, the less the departure from a 90° phase relationship between the reactive voltages E_2 and E_1, and also E_3 and E_1. The greater the frequency deviation, the greater is the difference in angular displacement between E_2 and E_1, and E_3 and E_1, so that the differential voltage obtained from the diodes is greater. When viewed from the angle of audio intensity, the greater the differential voltage from the rectifiers, the louder the audio signal, since the extent of deviation at the transmitter is a function of modulating voltage level. The greater the modulating voltage level within prescribed limits, the greater the frequency deviation.

In brief, then, the differential output voltage is a function of the rate of deviation of the f-m signal as well as the amount of frequency deviation. Since the *amplitude* of the audio modulating signal determines the amount of deviation and since the *frequency* of the audio determines the rate of change of the deviation of the f-m signal, it becomes readily apparent that the differential voltage across the two diodes will be an audio signal equal in frequency and proportional in amplitude to the audio modulation signal.

The Discriminator Output Curve

If the output voltage of the discriminator is plotted against the instantaneous values of frequency input, we obtain the well-known S-shaped response curve. A typical S-shaped discriminator response curve is illustrated in Fig. 5-15. The center point of the curve is at zero voltage output and at the center frequency of the f-m signal, namely the i.f. Above the zero point on the vertical axis, the output voltage is positive, and below the output voltage is negative. In similar fashion, to the right of zero on the horizontal axis the input f-m signal has instantaneous values of frequency greater than the i.f. and to the left the instantaneous frequencies will be lower than the i.f.

The actual shape of the curve is *not* determined by the frequency deviations of the input f-m signal. The shape of the S curve is determined by the design characteristics of the discriminator circuit. In order to have the output audio signal free from distortion, the input f-m signal to the discriminator should have frequency deviations that do *not* fall outside the linear range of the S characteristic. The linear range in the curve of Fig. 5-15 is indicated by points *A* and *B*. If the input f-m signal has frequency deviations that fall outside this linear range, they will operate in part over the nonlinear portion of the S curve and distortion in the audio output signal will result.

The curvature of this response characteristic is introduced by the nonlinearity of the detector outside the band of frequencies for which it is designed to operate.

We know that 50-kc peak-to-peak deviation is that for 100-percent modulation for f-m sound transmission for television, and therefore the linear portion of the S characteristic should be at least 50 kc wide. In fact, the greater the linearity (within reasonable limits) the better, and an over-all 200-kc linearity is considered very desirable. This accomplishes two things: First it means that the receiver does not have to be tuned very accurately for the resting frequency to fall in the middle of the discriminator characteristic. If the receiver is mistuned somewhat, no distortion will result provided the receiver is not so badly mistuned that the frequency variations in the signal extend into the nonlinear or curved portions of the discriminator characteristic. Secondly, the fact that the characteristic of the discriminator is linear over a greater range than that actually required means that the linearity will be more nearly perfect over the center portion which is actually used in reception. The high degree of linearity obtained in this way makes demodulation of the signal possible with practically no distortion.

Pre-emphasis and De-emphasis

In the audio input to a transmitter the amplitudes of the higher audio frequencies are relatively low as compared with the rest of the audio-frequency spectrum. This is due to the natural distribution of sound in radio program material, *not* to any characteristic of a transmitter. As the programs pass through the transmitter to the receiver and then through the receiver, noise is unavoidably added to the desired audio signal. This noise is predominantly high-frequency audio, and this condition tends to produce a low signal-to-noise ratio at the high audio frequencies because the signal is relatively weak and the noise relatively strong. In a communi-

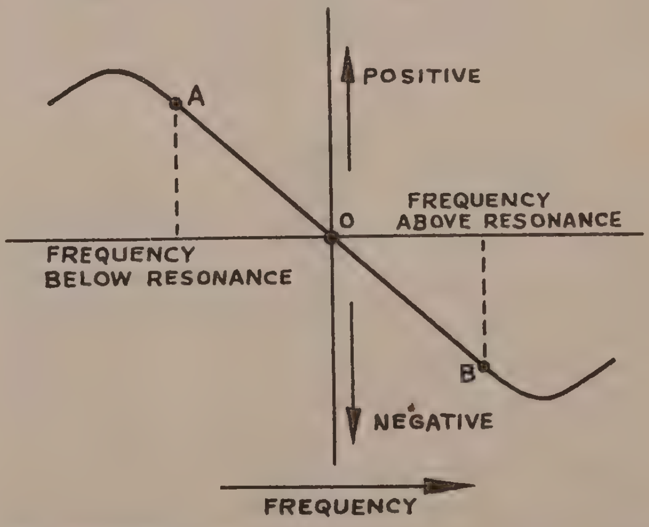


Fig. 5-15.—The S curve represents the typical output characteristic of the discriminator circuit with the output a-f voltage plotted against the instantaneous values of input frequency.

the voltage drop across R. The "effective bias curve" (idealized curve) represents the average bias. The true curve would be a bit irregular, because the diode current flows in pulses.

In practice, the values of R and C are so chosen that they represent a *time constant* somewhere in the vicinity of a few microseconds. This time constant means capacitor C will take so many seconds to lose about 63 percent of its charge through the resistor R. If C is, say 50 $\mu\mu f$ and R 50,000 ohms, then the time constant will be $R \times C = 50 \times 10^{-12} \times 50{,}000 = 2.5 \times 10^{-6}$ seconds or 2.5 microseconds. This time constant is considered high compared with the time it takes the input i-f signal to complete one cycle. With a 21-mc i-f signal, the time it will take the signal to complete one cycle is equal to 1/21 mc or about 0.05 microsecond. Comparing this value with the time constant of 2.5 microseconds, it is seen that one cycle of the 21-mc input signal will be completed 50 times as fast as that time represented by $R \times C$, and a half cycle of the input signal will be completed 100 times as fast. By the time one half cycle of input voltage is completed, the capacitor has discharged only about one percent of its voltage. Upon this very significant fact the operation of the limiter is primarily based, because it means that during the first negative half cycle not all of the voltage is discharged from C.

At the start of the second cycle of input signal, the grid voltage is different. A residual voltage remains on the capacitor, because the capacitor keeps a large portion of its charge during the negative half cycle of the signal input. Consequently, no grid current will be drawn during the positive half of the second cycle until this half cycle reaches a value whereby it will surmount the negative voltage remaining on the capacitor. This will occur at some time after the start of the positive half of the second cycle. During the negative half of the first cycle an operating bias was established due to the drop across R, and this bias voltage must be overcome by the positive half of the second cycle before grid current can be drawn. When this point is reached, grid current will start to flow again, and the capacitor begin to charge. Since the input signal is continually varying in a negative and positive direction, the capacitor is always being charged or discharged. Since grid current only flows during part of this positive half of the second input cycle, in which case capacitor C is being charged, then during the rest of the second cycle the capacitor will discharge through R, and a greater operating voltage will be developed than during the first cycle.

On the positive half of the third input cycle, a still greater bias has to be overcome before the grid will draw current. This means that a larger negative voltage on the capacitor has to be overcome before the capacitor can become charged. During the part of the cycle when the grid is not drawing current, the capacitor is discharging through R. Since a larger initial voltage existed across the capacitor at the beginning of this discharge, a still greater voltage drop appears across the resistance and, therefore, a higher operating bias. This process of a continually higher positive signal needed for grid current to flow for each succeeding cycle of input signal continues up to a certain point. Operating bias also increases with each successive input cycle until a constant bias is reached. The increase in signal input bias reaches a point where a *balance or equilibrium* occurs in the system. After a certain number of input cycles the grid will reach a point where constant *operating bias* on the tube, and the charge and discharge of the capacitor during each succeeding cycle after this point has been reached, will remain the same. The applied d-c voltages on the tube elements, the RC time constant, and the strength of the input signal determine this balance point.

Coming back to Fig. 5-6, the operating bias on this i_b-e_c curve is seen to be balanced at —6 volts. This means that after equilibrium is reached the input signal will swing about the operating bias of —6 volts. A glance at the curve will show that —8 volts on the grid (whether fixed or instantaneous) will drive the plate current to cutoff. Thus the negative half of the input signal need swing only 2 volts to result in plate current cutoff. Consequently, the rest of the negative cycle is clipped, or cut off in the plate circuit. This clipping action of the negative peak is shown in the diagram. Clipping of the positive peaks as previously discussed takes place in the grid circuit as a result of the diode action. On these positive peaks the operating bias controls grid current flow to a point where the grid is barely driven positive. This is readily noticed in the diagram where more than 6 volts swing on the positive half cycle is needed to drive the grid into the positive region for grid current flow. In this manner the grid acts as a diode rectifier in conjunction with the cathode, and clipping of the positive peaks results.

The time constant indicates how quickly the grid bias will change with change in level of the input signal to produce amplitude limitation. The amplitude variations in the final output i-f signal due to the response characteristics of the i-f circuit are not radical or sudden changes but gradual ones. As a result the time constant of the network need not be very low to limit such amplitude variations, and time constants of 10 microseconds would be sufficient. But one finds that in practice time constants between 1 and 5 microseconds are commonly used so that the operating bias can follow any quick rise in amplitude of the input signal resulting from sudden noise impulses or other interference. If the time constant were too high for the bias to follow this impulse, plate current would flow in accordance with the impulse variation, and the noise impulses would be passed on to the discriminator to be detected and, eventually, reproduced by the audio system.

AVC From Limiters

In a-m receivers when automatic volume control (avc) is employed it is taken from the diode detector. When avc is used in the sound channel of a television receiver employing the limiter-discriminator method of detection, it is often taken from the grid circuit of the limiter. It will be recalled that due to the discharging of the capacitor of the time constant network, a bias is developed across the grid-leak resistor. Also, the stronger the input signal, the greater will be the bias developed across the resistor. Another glance at Fig. 5-5 will reveal that, due to the current flow through R, the grid side of the resistor is made negative with respect to ground. Since the bias on this tube is a function of the instantaneous input signal and becomes more negative with increase in signal input, we have a point from which avc voltage can be obtained.

Input Voltage Considerations

One of the rudiments of correctly using a limiter system is to know the minimum amount of input signal required to have the system limit properly and, in addition, produce an output signal strong enough for detection. As explained before, a certain threshold voltage has to be met to have the stage following the last i-f stage act as a limiter. Therefore, the output of the last i-f stage must be of sufficient strength to drive the limiter tube beyond the threshold point of operation. Since the last i-f stage has to drive the limiter tube, all the previous stages together must have produced enough signal output to cause the last i-f tube to drive the limiter tube properly. Since the input signal to the receiver times the *fixed* amount of gain of the r-f and i-f stages determines the signal input to the limiter system, it is evident that, for a given receiver, the variable factor of signal pickup determines whether or not there will be enough voltage to drive the limiter.

From the foregoing factors we see that for the sound channel of a television receiver with a fixed amount of r-f and i-f gain, the f-m input signal from the antenna to the first stage of the television receiver has to be above a fixed value. For instance, if the required threshold voltage of the limiter tube is 2.5. volts and if the over-all r-f and i-f gain is equal to 100,000 times, the necessary f-m input signal to the first stage of the receiver is 2.5 volts/100,000 or 25 microvolts. This means that any f-m signal input to the receiver below 25 microvolts will not drive the limiter sufficiently to cause limitation.

From this analysis it is seen that the limiter stage has a very decided influence on the design of the other stages and definitely influences the gain, sensitivity, and selectivity of the receiver.

The selectivity of the i-f stages helps determine the necessary amount of f-m input signal to the antenna. This is best explained in terms of the i.f.'s response characteristics. The amplitude of a signal passed through the i-f network is not constant for the complete range of frequencies the network is designed to pass. Since all these frequencies must be limited, that part of the i-f curve that produces the lowest amplitude must be the criterion in the design of the gain of the receiver. In other words, with the proper r-f and i-f gain, this i-f band-pass signal must exceed the threshold voltage required for the limiting tube at the smallest amplitude within its band.

As mentioned, the limiter tube has a certain amount of amplification, but beyond some value of signal input the plate current of the limiter tube will not increase but will remain constant and a constant output voltage is the result.

THE DISCRIMINATOR SYSTEM

Immediately following the limiter is the discriminator stage. The many different types of discriminator networks all have the primary function of demodulating the incoming f-m signal. It should be remembered that this input f-m signal is constant in amplitude due to the action of the preceding limiter stage. The type of discriminator circuit used in the sound channel of television receivers is similar to that used in f-m receivers. The general form of this circuit appears in Fig. 5-8. There are some slight modifications of this circuit which appear in certain sound channels of television receivers and these will be discussed later on. This type of circuit, known as the Foster-Seeley discriminator, is also known as the phase discriminator, the center-tapped secondary discriminator or the center-tuned type of discriminator. In many instances this type of circuit is referred to just as the discriminator.

The purpose of the discriminator is to demodulate the f-m input signal so that in its output circuit will appear an audio signal that is varying in frequency in accordance with the audio modulation of the f-m signal and is proportional in amplitude to the amplitude of this audio modulation.

Circuit Analysis

From Fig. 5-8 we notice that the discriminator consists of two diodes which can be separate tubes but are usually a single duo-diode tube. Associated with the discriminator is the tuned discriminator transformer whose primary coil L_1 is in the plate circuit of the limiter tube and the center-tapped secondary coil L_2 is connected to the plates of the discriminator. The primary tuned circuit L_1C_1 and the secondary tuned circuit consisting of inductance $L_2 + L_3$ and capacitor C_2 are both resonant to the i.f.

The two diodes are connected in the form of a differential rectifier system where the two resistors R_1 and R_2 are the respective loads for diodes D_1 and D_2. The d-c paths through these diode circuits are completed through coil L and the respective half of the center-tapped coil associated with each circuit. Besides providing the d-c paths between the diode plates and their respective cathodes, this common coil has another function which will be shown later. In some discriminator circuits the coil L is replaced by a resistor and in others no coil or resistor is used. However, these changes do not change the operation of the discriminator circuit. This will be seen later when modifications of the discriminator circuit are analyzed. Let us now discuss the discriminator circuit of Fig. 5-8 with respect to the voltages illustrated therein.

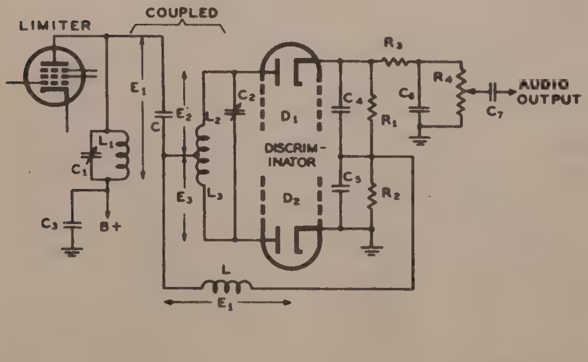


Fig. 5-8.—Schematic of a conventional phase discriminator detector.

Voltage E_1 is the i-f signal voltage developed across the tuned primary circuit. Examining the secondary of this i-f transformer, we note certain significant details. It consists of two windings L_2 and L_3 in series, resonated to the i-f peak by means of C_2. The center tap on the secondary winding is connected to a coupling capacitor C and, also, to an r-f choke L.

Associated with the two circuits and the r-f choke L are three voltages, designated as E_2, E_3, and E_1 respectively, the latter being virtually identical to E_1 across the i-f transformer primary. To explain these designations, it is necessary to discuss the coupling between the primary and secondary circuits of this transformer, as well as what happens in a transformer when the secondary is tapped at the midpoint. How does this type of discriminator operate? In brief the operation can be divided into three major actions, although more conditions than just three are actually involved.

In the first place, although a single tuned winding is used for the secondary circuit, the center tap on this winding causes a division of the signal voltage developed in the tuned circuit across the two halves of the secondary winding, that is across L_2 and L_3. The signal voltages across these two halves are always equal to each other, irrespective of the frequency of the signal voltage fed into this circuit from the primary.

The second major consideration is that the signal voltage present across the primary winding L_1 is also present across winding L, which is common to both halves of the secondary circuit with respect to the signal voltages eventually applied to the two diodes D_1 and D_2.

The final major action is the phase relation which exists between the signal voltage across L_2, which we can call E_2, and the signal voltage across L which, because it is the same as that across L_1, is also identified as E_1; also the phase relation between the signal voltage across L_3, or E_3, and the signal voltage across L, or E_1. The function of this discriminator network with particular reference to these three actions will now be discussed in detail.

Two methods of coupling the signal from the primary to the secondary circuit are used in this system. The resonant primary is inductively coupled to the resonant secondary winding; at the same time the signal voltage E_1 across the primary is fed to the r-f winding L via the coupling capacitor C. If the circuit of C, L, and C_5 is traced, it will be seen that L is in shunt with the tuned primary grounded through C_3. None of the quantities C, L, C_3 or C_5 is of a magnitude to alter the resonant conditions of C_1 and L_1, the resonant primary. Thus, with respect to magnitude and phase, whatever signal voltage exists across C_1-L_1, also exists across L. The direct connection between the coupling capacitor C and the mid-point of the secondary winding is of no consequence to the signal transfer between the primary and the secondary tuned circuits; however, it is the point to which the choke L must be connected to complete the differential rectifier circuit. Thus, the secondary system receives signal voltages in two ways: the resonant secondary receives its signal voltage by inductive coupling, and the r-f choke derives its signal voltage by means of capacitive coupling through the fixed capacitor C.

The equal voltages across each half of the secondary winding are obtained in the following manner. When a winding is tapped at the mid-point and a voltage is induced in that winding by means of a varying magnetic field, the total voltage developed across the entire winding divides between the two halves. This is logical in view of the fact that half the total number of turns exists between the center tap and one end, and half between the center tap and the other end. So, whatever the nature of the signal voltage which will be developed across the tuned secondary circuit C_2-L_2-L_3, it is possible to show that this voltage divides into two parts, that is, across each half of the winding. These voltages are designated as E_2 and E_3.

The above analysis of the discriminator circuit will enable us to understand the discussion that follows. We will see how the input f-m signal to the discriminator becomes demodulated resulting in an audio signal at the output which is the audio modulation of the input f-m signal. In order to show how this occurs we have to consider the input f-m signal (which is varying in frequency above and below its center value) at three different instants of time. We will investigate how the circuit functions at the time when the f-m signal is exactly equal to the tuned i.f. of the discriminator transformer (i.e. at resonance) and at the time when its instantaneous frequency is above and below the center i.f. Phase relationships among the various voltages are the criterion in establishing the operation of the discriminator. That is one of the reasons why the circuit is often known as the phase discriminator.

Resonance Conditions in the Phase Discriminator

Let us first examine this circuit at resonance. At resonance, the frequency of the applied signal and the resonant frequencies of the tuned circuits are both the same. Since the inductances and capacitances of a tuned circuit effectively cancel each other at resonance, the circuit behaves like a resistance. In a resistive circuit the current is in phase with the voltage, so in the secondary tuned circuit the induced current, call it I, caused to flow by the induced voltage, call it E, is in phase with this induced voltage. It should be remembered that this induced voltage is effectively *in series* with the inductance and capacitance of the secondary tuned circuit.

The in-phase relationships between E and I are indicated in the vector diagram of Fig. 5-9 where vectors OI and OE, the respective induced voltage and current vectors, are seen to be in phase with each other. The voltage across the primary circuit, designated as E_1 is 180° out of phase with the voltage induced into the secondary circuit. This voltage E_1 is the main voltage upon which all the other voltages are based. Consequently, this voltage is drawn as vector OE_1 along the 0° reference line.

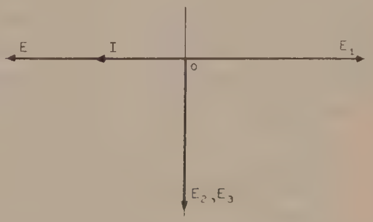


Fig. 5-9.—Vector diagram of the relations at resonance between the current and voltages in the phase discriminator.

In the vector diagram of Fig. 5-9, vector OE_1 is 180° out of phase with the induced voltage vector OE. Since the voltage across a pure inductance leads the current through it by 90°, the voltage drops E_2 and E_3 across the secondary coil (called reactive voltage drops because the inductance is considered to be a pure inductance containing negligible resistance) lead the current I flowing through it. This is indicated where vectors OE_2 and OE_3 are leading the induced current by 90°. This also means that the reactive voltage drop across the secondary coil is lagging the primary voltage E_1 by 90°, and vector OE_1 is seen to be leading vectors OE_2 and OE_3 by 90°. The 90° phase difference between these voltages is very important to the operation of the discriminator. It should still be remembered that this voltage E_1 also exists across coil L of Fig. 5-8 in the same phase and magnitude as that existing across the primary tuned circuit.

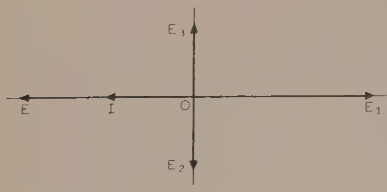


Fig. 5-10. — Revised vector diagram of Fig. 5-9. Note that the vector OE₃ is shifted 180° in phase from its position in the previous figure.

That the secondary is center tapped means that it is in a push-pull arrangement, and hence voltages E_2 and E_3 are equal in magnitude but 180° out of phase with each other, as referred to the center tap. However, the same current flows through both parts of the secondary coil, so that a 90° phase relation must still exist between each voltage and the current; but in one case one of the voltages is effectively leading the current and in the other case the voltage is effectively lagging the current, by 90°. This means too that one half of the secondary voltage drop is leading voltage E_1 by 90° and the other half lagging voltage E_1 by 90°. All of this is indicated in the revised vector diagram of Fig. 5-9 as shown in Fig. 5-10, in which vector OE_3 of the previous vector diagram has been shifted 180°.

To demonstrate how all these voltages affect the duo-diode circuit we have redrawn that part of Fig. 5-8 appearing to the right of the secondary of the transformer in simple form in Fig. 5-11. In this figure we have made two separate circuits of the diodes, showing the respective voltages that act upon each diode. These two simple circuits then are combined to show how they actually work together. This figure and Fig. 5-8 show that voltage E_1 is common to both diodes since it exists across the inductance L. Also, since voltage E_2 is active on diode D_1 and voltage E_3 is active on diode D_2, it is readily seen from Fig. 5-11 that voltages E_2 and E_1 are active on diode D_1 and voltages E_3 and E_1 are active on diode D_2.

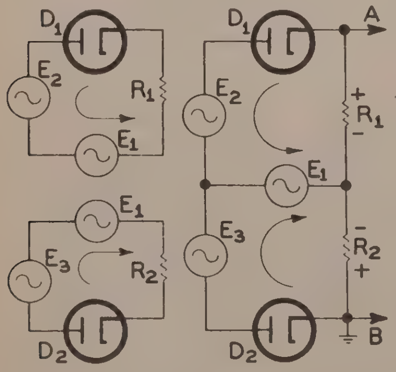


Fig. 5-11. — Simplified schematics of the diodes of Fig. 5-8 and their respective load resistors. These two schematics are combined on the right. This is permissible since the the voltage E₁ is common to both diodes.

Further examination of this simplified circuit reveals that the rectified current flowing through the individual diode circuits puts certain polarities on their load resistors. Since the external current in a diode rectifier circuit flows from plate to cathode, the currents in the diode load resistors will be flowing in opposite directions, and the polarities across the individual load resistors will be bucking each other. Thus between points A and B a voltage will exist which will be the difference between the voltage drops across resistors R_1 and R_2.

If the voltages E_2 and E_3 have the *same phase angle* with respect to the voltage E_1, both diode currents will be equal in value, and the same voltage drop will appear across each load resistor R_1 and R_2. Since each resistance voltage drop is opposite in polarity to the other but equal in value, the total voltage measured between points A and B will be zero. Under these circumstances the output of the differential rectifier circuit is zero. If, however, the phase relationships between E_2 and E_1 differ, a differential voltage will exist between points A and B, because the voltage drops across R_1 and R_2 no longer will be equal to each other as different currents flow through each diode.

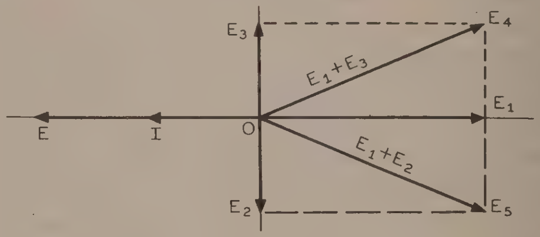


Fig. 5-12.—Vector diagram of the current and voltage relationships of the phase discriminator of Fig. 5-8 at resonance when both secondary voltages E₂ and E₃ have a 90° phase difference from E₁.

Zero voltage exists across points A and B when the resonant frequency of the tuned discriminator i-f transformer is exactly equal to the applied frequency. This is simply indicated by the vector diagram of Fig. 5-12 where both secondary voltages have the same phase difference, namely 90°, from voltage E_1. This diagram is nothing more than a duplicate of that diagram of Fig. 5-10 with the exception that the two voltages active on each diode are added *vectorially*. Thus, in Fig. 5-12, vector OE_4 represents the resultant vector of the vector addition of voltages E_3 and E_1 across diode D_2 and vector OE_5 represents the resultant vector of the vector addition of E_2 and E_1 across diode D_1. Resultant vectors OE_4 and OE_5 are shown to be equal in magnitude, causing the same current to flow in each diode circuit. Thus, equal but opposite voltages are developed across diode load resistors R_1 and R_2 producing zero voltage between points A and B.

Applied Frequency Higher Than Resonance

When the instantaneous value of the f-m signal is equal to its center frequency component, we have the frequency applied to the discriminator transformer equaling the resonant frequency. The situation for this was discussed in the preceding section. At either side of the center frequency component of the f-m signal the instantaneous frequency is different from the resonant frequency of the i-f transformers. Under these conditions the discriminator transformer is tuned below or above the incoming i-f signal.

Let us now consider an instantaneous value of the f-m signal greater than the center i.f. The discriminator transformer then is tuned below the incoming i.f. The circuit is still the same as in Fig. 5-8, and the nonresonant conditions do not alter the fundamental rules of the action of the primary circuit, so that the voltage E_1 that exists across this circuit also exists across L in both the same phase and magnitude.

Induced voltage E in the secondary remains 180° out of phase with the primary signal, for this too is a fundamental condition which is not altered by nonresonance conditions. However, the phase relationship betwen the induced voltage E and the current I, which it causes to appear in the secondary circuit, is affected by the state of resonance, and in turn alters related conditions.

When the applied frequency is higher than the resonant frequency, the reactance of the secondary coil becomes greater than the reactance of the capacitor. This accords with the fundamental law that inductive reactance varies directly with frequency, and capacitive reactance varies inversely with frequency. Accordingly, a portion of the inductive reactance will be offset by the capacitive reactance, but a certain amount of inductive reactance will remain to exert a control on the induced current. The circuit as a whole now appears as an inductance and resistance in series, rather than as a resistance alone, which is the case at resonance.

Under this circumstance the induced current I no longer will be in phase with the induced voltage E but rather will lag this voltage by a certain amount, depending upon the extent to which the instantaneous f-m signal is greater than the tuned frequency of the transformer. This is all indicated in the vector diagram of Fig. 5-13 for the off-resonance condition now being discussed. Voltages E and E_1 are still seen to be 180° out of phase but the phase relationships of the other component voltages differ somewhat from those of the vector diagram of Fig. 5-12.

For the sake of argument let us say that the difference in frequency between the instantaneous frequency of the f-m signal and the tuned i-f transformer is such that the amount of inductive reactance remaining is sufficient to cause the induced current I to lag the induced voltage E by 35°.

No matter what the phase relationship between the induced voltage and induced current, the two voltages E_2 and E_3 across the individual halves of the secondary are still 180° out of phase with each other and equal in magnitude. The induced current flowing through this secondary still bears the same phase relationship to these secondary voltages. Regardless of the phase difference between E and I, secondary voltage E_3 will still lag current I by 90°, and secondary voltage E_2 will still lead current I by 90°. This is indicated in the vector diagram of Fig. 5-13; and if this vector diagram and that of Fig. 5-12 are compared, these phase relations will be seen to hold.

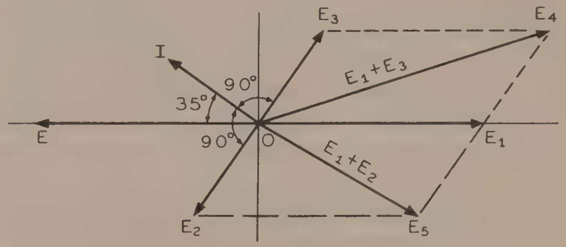


Fig. 5-13.—Vector diagram of the current and voltage relationships in the phase discriminator when the instantaneous frequency of the signal is higher than the resonant frequency of the transformer.

Let us further compare these two vector diagrams. To keep constant the 90° phase relations between voltages E_2, E_3, and current I, then when current I lags induced voltage E by 35°, voltage vectors OE_2 and OE_3 are both shifted 35° clockwise to keep these 90° relationships intact. The complete vector line E_2-O-E_3 is shifted 35° in a negative direction. When the respective voltages applied to the individual diodes are added under these circumstances, it will be seen from the vector diagram of Fig. 5-13 that the resultant vector OE_4 representing that voltage across diode D_2 and resultant vector OE_5 representing that across diode D_1 are no longer equal, but that vector OE_4 is greater than vector OE_5. In this instance diode D_2 will draw the greater current, and in Fig. 5-11 load resistor R_2 will have a greater voltage drop than resistor R_1; hence a differential voltage will exist across point *A* to *B*, with point *B* being more positive than point *A*. This is the same as saying point *A* is negative with respect to point *B*.

Applied Frequency Lower Than Resonance

When the instantaneous value of the f-m signal input to the discriminator circuit is such that it is less than the resonant frequency of the discriminator transformer, the differential voltage will still exist across the diode loads, but the polarities will be reversed. Let us see how this happens.

We still are at off-resonance conditions, even though we are on the lower side of the resonant frequency, and the same 180° phase relationship between *E* and E_1 exists. When the applied frequency is lower than that of the resonant frequency of the i-f transformer, the impedance of the secondary of the i-f transformer is such that the capacitive reactance more than balances out the inductive reactance, and the secondary is primarily capacitive. Since this circuit is capacitive, the induced current *I* leads the induced voltage *E*. If the off-resonance conditions are such that a phase angle of 35° again exists between *I* and *E*, *I* will be *leading* *E* by 35°, as seen in the vector diagram of Fig. 5-14. Since the 90° phase relations between voltage E_2 and current *I*, and voltage E_3 and current *I* must still exist, these two voltages are effectively shifted in phase 35° in a counterclockwise or positive direction. This is indicated in Fig. 5-14 where vectors OE_2 and OE_3 are still 180° out of phase with each other, but no longer 90° out of phase with vector OE_1. Now when the individual i-f voltages across the diode circuits are combined vectorially, it will be seen that resultant vector OE_5 applied to diode D_1 is greater in magnitude than resultant vector OE_4 applied to diode D_2. Therefore, the current in the circuit of diode D_1 is greater than the other diode current. This means a greater voltage drop exists across R_1, the load resistor of diode D_1, than across R_2, the load of D_2, and a differential voltage exists between points *A* and *B* of the diode circuit of Fig. 5-11. However, under these conditions the polarity of point *A* will be more positive with respect to the junction of R_1 and R_2 than point *B*. That is, point *B* is negative with respect to point *A*.

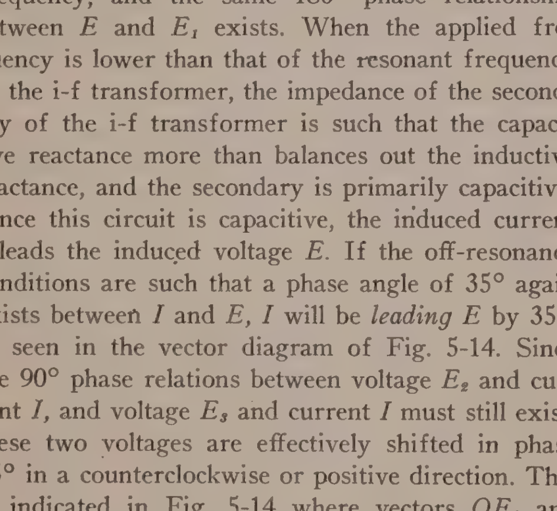


Fig. 5-14.—When the instantaneous frequency of the f-m signal is lower than the resonant frequency of the transformer, the current I will lead the induced voltage E. Compare with Figs. 5-12 and 5-13.

Summary

Summarizing the action described, it is evident that if a varying frequency input signal (one which varies in frequency around a mean) is applied to the discriminator network—provided that the range of frequencies covered is not beyond the acceptance bandwidth of the discriminator transformer—an output signal which changes in amplitude and polarity will be obtained. The frequency swing of the f-m signal is determined by the amplitude of the audio modulating signal. The greater the amplitude, the wider the frequency swing, and the smaller the amplitude, the lesser the frequency swing. The frequency of the audio modulating signal determines the time rate change of the frequency swing. This frequency swing, usually called *frequency deviation*, varies an equal amount on either side of the center or carrier frequency of the f-m signal for each cycle of audio modulating signal. If the audio modulation is a sine wave of constant amplitude or any similar signal of constant amplitude, the frequency deviation will be the same for every cycle of modulation. However, in speech or music the audio modulation is anything but constant in amplitude and, therefore, the frequency deviation for each cycle is usually different.

The output signal from the discriminator is determined by the amount of frequency deviation; the less the frequency deviation, the less the departure from a 90° phase relationship between the reactive voltages E_2 and E_1, and also E_3 and E_1. The greater the frequency deviation, the greater is the difference in angular displacement between E_2 and E_1, and E_3 and E_1, so that the differential voltage obtained from the diodes is greater. When viewed from the angle of audio intensity, the greater the differential voltage from the rectifiers, the louder the audio signal, since the extent of deviation at the transmitter is a function of modulating voltage level. The greater the modulating voltage level within prescribed limits, the greater the frequency deviation.

In brief, then, the differential output voltage is a function of the rate of deviation of the f-m signal as well as the amount of frequency deviation. Since the *amplitude* of the audio modulating signal determines the amount of deviation and since the *frequency* of the audio determines the rate of change of the deviation of the f-m signal, it becomes readily apparent that the differential voltage across the two diodes will be an audio signal equal in frequency and proportional in amplitude to the audio modulation signal.

The Discriminator Output Curve

If the output voltage of the discriminator is plotted against the instantaneous values of frequency input, we obtain the well-known S-shaped response curve. A typical S-shaped discriminator response curve is illustrated in Fig. 5-15. The center point of the curve is at zero voltage output and at the center frequency of the f-m signal, namely the i.f. Above the zero point on the vertical axis, the output voltage is positive, and below the output voltage is negative. In similar fashion, to the right of zero on the horizontal axis the input f-m signal has instantaneous values of frequency greater than the i.f. and to the left the instantaneous frequencies will be lower than the i.f.

The actual shape of the curve is *not* determined by the frequency deviations of the input f-m signal. The shape of the S curve is determined by the design characteristics of the discriminator circuit. In order to have the output audio signal free from distortion, the input f-m signal to the discriminator should have frequency deviations that do *not* fall outside the linear range of the S characteristic. The linear range in the curve of Fig. 5-15 is indicated by points *A* and *B*. If the input f-m signal has frequency deviations that fall outside this linear range, they will operate in part over the nonlinear portion of the S curve and distortion in the audio output signal will result.

The curvature of this response characteristic is introduced by the nonlinearity of the detector outside the band of frequencies for which it is designed to operate.

We know that 50-kc peak-to-peak deviation is that for 100-percent modulation for f-m sound transmission for television, and therefore the linear portion of the S characteristic should be at least 50 kc wide. In fact, the greater the linearity (within reasonable limits) the better, and an over-all 200-kc linearity is considered very desirable. This accomplishes two things: First it means that the receiver does not have to be tuned very accurately for the resting frequency to fall in the middle of the discriminator characteristic. If the receiver is mistuned somewhat, no distortion will result provided the receiver is not so badly mistuned that the frequency variations in the signal extend into the nonlinear or curved portions of the discriminator characteristic. Secondly, the fact that the characteristic of the discriminator is linear over a greater range than that actually required means that the linearity will be more nearly perfect over the center portion which is actually used in reception. The high degree of linearity obtained in this way makes demodulation of the signal possible with practically no distortion.

Pre-emphasis and De-emphasis

In the audio input to a transmitter the amplitudes of the higher audio frequencies are relatively low as compared with the rest of the audio-frequency spectrum. This is due to the natural distribution of sound in radio program material, *not* to any characteristic of a transmitter. As the programs pass through the transmitter to the receiver and then through the receiver, noise is unavoidably added to the desired audio signal. This noise is predominantly high-frequency audio, and this condition tends to produce a low signal-to-noise ratio at the high audio frequencies because the signal is relatively weak and the noise relatively strong. In a communi-

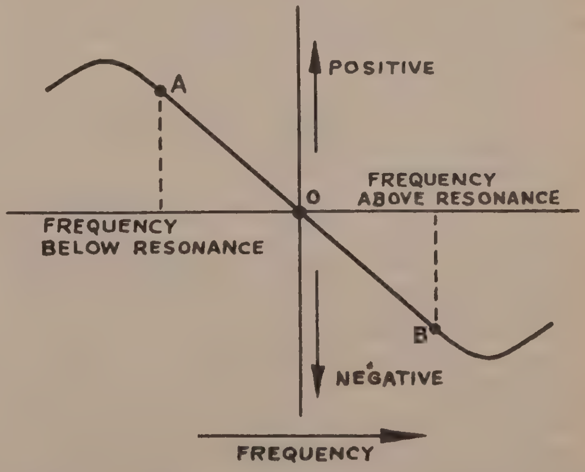


Fig. 5-15.—The S curve represents the typical output characteristic of the discriminator circuit with the output a-f voltage plotted against the instantaneous values of input frequency.

cations system, such as commercial f-m broadcasting, where the audible spectrum up to 15,000 cycles is utilized, this low signal-to-noise ratio at the high audio frequencies is most undesirable. However, it is possible to overcome this condition to a large extent by the use of *pre-emphasis* which means increasing the relative strength of the high-frequency components of the audio signal. However, at the same time that pre-emphasis is beneficial in improving the signal-to-noise ratio, it introduces a defect in that it upsets the natural balance between the high- and low-frequency tones in the program material.

This defect is compensated for at the television receiver by means of a *de-emphasis* circuit at the input to the audio amplifier. The de-emphasis circuit *reduces* the high-frequency audio exactly as the pre-emphasis *increases* it. However, it operates on *both* the high-frequency program material *and* the high-frequency noise. Thus, it does not change the *improved* high-frequency audio signal-to-noise ratio, which is obtained by means of pre-emphasis, while it does re-establish the tonal balance of the program material, which is lost in pre-emphasis.

This de-emphasis network is usually inserted between the detector and audio amplifier in the sound channel and has a frequency characteristic just opposite to that of the pre-emphasis network.

The FCC has set a standard of a 75-microsecond time constant for the pre-emphasis network in the transmitter. Consequently, the de-emphasis network in the receiver also should have a time constant equal to 75 microseconds.

A typical de-emphasis network is shown in Fig. 5-16. The effective impedance offered to the audio voltage is the series combination of C and R. As the frequency of the audio signal increases, the reactance of the capacitor C decreases, or as the frequency increases, the reactive voltage drop decreases. The audio voltage to the grid of the tube, therefore, decreases with increase in frequency and the reverse of pre-emphasis is obtained. The time constant is determined similarly to the method described in the discussion of limiters, in that it is equal to the product of R in ohms by C in farads.

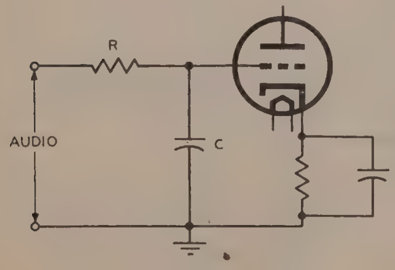


Fig. 5-16.—A typical de-emphasis network as used in the sound section of television receivers.

In Fig. 5-8 the R_3 and C_6 combination is a de-emphasis that appears in the output of the detector network before the audio signal is applied to the preceding audio amplifiers.

De-emphasis networks also appear in the output of the ratio detector as well as the discriminator.

Modification of the Discriminator Network

We mentioned awhile back that there exist modifications of the discriminator network of Fig. 5-8. These modifications vary somewhat but their functions are all the same. That is, the voltage existing across the primary of the discriminator transformer essentially exists across each diode. In some of the modified circuits the method of coupling this voltage is readily noticed, while in others it is not.

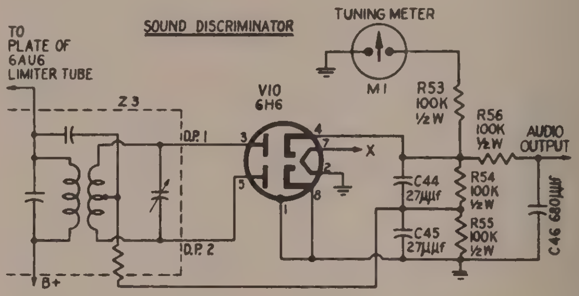


Fig. 5-17.—Sound discriminator circuit of the Du Mont model RA-102. Note that the coil L in Fig. 5-8 has been replaced by resistor R.

Courtesy Allen B. Du Mont Laboratories, Inc.

Most of the discriminator circuits of today as used in television sound channels do not employ a coil (such as L in Fig. 5-8) for directly coupling the signal voltage across the discriminator primary to the two diodes. The modified discriminator circuits discussed below are the more common types found in use today.

The sound discriminator circuit of the Du Mont television receiver model RA-102 is reproduced in Fig. 5-17. The only difference between this circuit and that of Fig. 5-8 is that coil L in the latter is replaced by resistor R in the former. This resistor R performs the same function as coil L in that the signal voltage drop across the primary of the discriminator transformer also appears across resistor R effectively in the same phase and magnitude as across coil L. The rest of the discriminator circuit of Fig. 5-17 appears and functions the same as that of Fig. 5-8. The tuning meter *MI* in Fig. 5-17 is used as an indicating device to show when the station is in tune. When properly tuned, the needle of this meter is oriented in the center of the scale. The *R*56 and *C*56 combination represents the de-emphasis network.

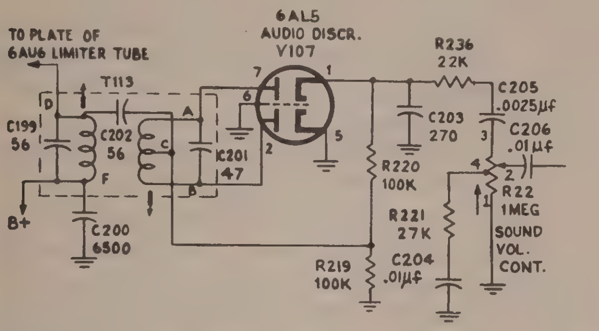

6AL5
AUDIO DISCR.
V107
TO PLATE OF
6AU6 LIMITER TUBE
T113
D
C199
56
C202
56
A
C
C201
47
B
F
B+
C200
6500
7
6
1
2
5
R236
22K
C205
.0025μf
C203
270
3
C206
.01μf
4
2
1
R220
100K
R22
1MEG
SOUND
VOL.
CONT.
R221
27K
C204
.01μf
R219
100K

Fig. 5-18.—Discriminator detector used in the Crosley models 307TA, 307TA-50.
Courtesy Crosley Div. AVCO Mfg. Corp.

Another modification of the discriminator appears in Crosley models 307TA, 307TA-50. Comparing the circuit of Fig. 5-18 with that of Fig. 5-8, we find that coil *L* is missing and that only one capacitor, *C*203, appears across the discriminator load resistors. It may not be readily apparent how the signal voltage across the discriminator primary of Fig. 5-18 is coupled to both diodes, although the transformer-coupled voltages that exist across each individual half of the discriminator secondary of Fig. 5-18 effectively appear across their respective diodes in similar fashion to that discussed for Fig. 5-8. However, it may not be readily evident how the signal voltage across the discriminator primary is directly coupled to both diodes, since coil *L* of Fig. 5-8 or resistor *R* of Fig. 5-17 is not present.

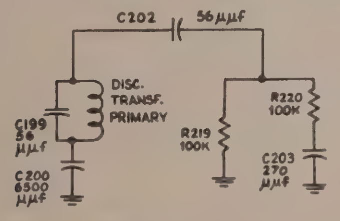

C202
56μμf
DISC.
TRANSF.
PRIMARY
C199
56
μμf
C200
6500
μμf
R219
100K
R220
100K
C203
270
μμf

Fig. 5-19. — The two paths of Fig. 5-18 from the top of the discriminator primary coil to ground are redrawn for the sake of the discussion.

If we trace the circuit of Fig. 5-18 from the top of the discriminator primary coil to ground. we can take two paths after passing through coupling capacitor *C*202. These two paths are redrawn in Fig. 5-19. The output voltage from the limiter is effectively across the discriminator transformer primary and capacitor *C*200 to ground. The 6500 *μμf* value of *C*200 has a reactance a little over one ohm at 20 mc, so practically all of the limiter output voltage appears across the primary coil. Across this coil and *C*200 capacitor arrangement is capacitor *C*202 and the two parallel resistor networks as shown in Fig. 5-19. Resistor *R*219 is in parallel with resistor *R*220 and capacitor *C*203. All the components to the right of the discriminator primary have an amount of the i-f signal voltage across them, proportional to their respective reactance or resistance. However, at an i.f. of about 20 mc, the reactance of capacitors *C*202 and *C*203 is very small (that of *C*202 being about 150 ohms and that of *C*203 being about 30 ohms) so that practically all the signal voltage appearing across the discriminator primary also appears across both load resistors *R*219 and *R*220. Since resistor *R*219 is the load for one diode and resistor *R*220 the load for the other diode, it is readily seen how the signal voltage which appears across the discriminator primary also appears across each diode.

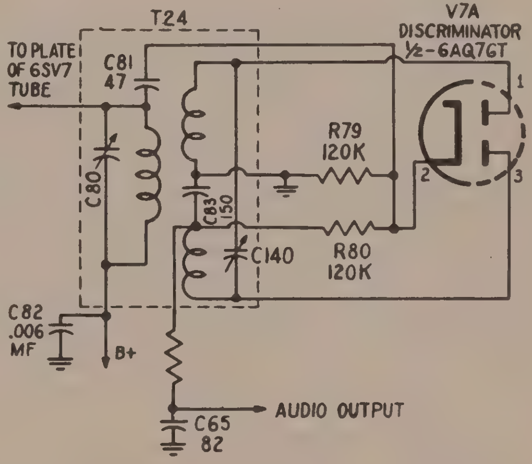

T24
V7A
DISCRIMINATOR
½-6AQ7GT
TO PLATE
OF 6SV7
TUBE
C81
47
C80
C83
150
R79
120K
C140
R80
120K
1
2
3
C82
.006
MF
B+
AUDIO OUTPUT
C65
82

Fig. 5-20.—The discriminator circuit used in the General Electric model 802.
Courtesy General Electric Co.

Another type of modified discriminator uses the diode sections of a duo-diode high-mu triode. The 6AQ7-GT, as used in the General Electric model 802, is such a tube. The way the discriminator circuit appears in the over-all schematic is illustrated in Fig. 5-20. The diodes of this tube can be used as a discriminator, because one cathode is used for the diodes and a separate one for the triode section of the tube. A schematic of the 6AQ7-GT tube appears in Fig. 5-21. From this drawing and that of

Fig. 5-20 we can see how the two diode sections are utilized. In order to understand better how this circuit functions we have redrawn it in Fig. 5-22. The voltages, coils, diodes, and load resistors are given designations to conform to those used in the conventional discriminator circuit of Fig. 5-8 so that we can compare them.

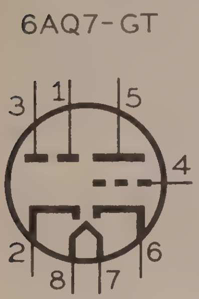


Fig. 5-21 (left) — Schematic of the 6AQ7-GT duo-diode high-mu triode tube.

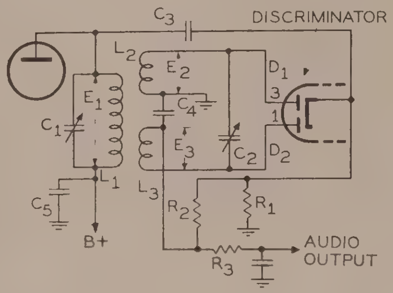


Fig. 5-22 (right) —The discriminator circuit of Fig. 5-20 has been redrawn to show the use of the 6AQ7-GT more clearly.

The chief difference between the circuit of Fig. 5-22 and the conventional discriminator is the method of applying reference voltage E_1 to both diodes D_1 and D_2. Coil L_2, and thus voltage E_2, are common to the upper diode D_1, and coil L_3 and voltage E_3 are common to the lower diode D_2. Capacitor C_4 connecting L_2 and L_3 is of high enough capacitance, so that both coils are effectively in series to the i.f. As with the conventional discriminator, capacitor C_2 is shunted across these two coils and with them forms the secondary tuned circuit. This analysis reveals how the respective induced voltages E_2 and E_3 are applied across the individual diodes D_1 and D_2, but the method of obtaining E_1 across both diodes is not readily evident.

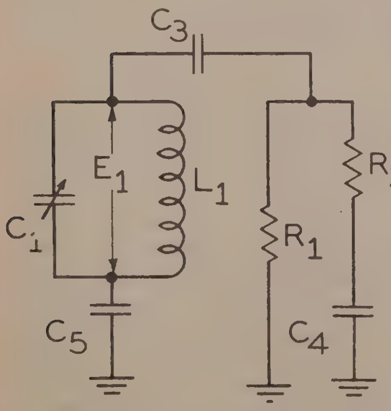


Fig. 5-23.—A simplified drawing of the i-f paths that appear across the transformer primary to ground in Fig. 5-22.

Tracing the d-c path for each diode, we find that resistor R_1 is the load for D_1 and resistor R_2 is the load for D_2. The reference voltage E_1 is capacitance coupled through C_3 to the common cathode of the diodes. Capacitor C_5 is a bypass capacitor for the primary tuned circuit and completes the r-f path to ground. The reactance of C_5 at the i.f. is so small that negligible i-f voltage appears across it.

If we trace the i-f path from the top of L_1, we find that there are essentially two parallel paths which appear across the primary circuit to ground. This circuit is shown in simplified form in Fig. 5-23. Going from the top of L_1 we pass through C_3, and find two paths available: one through resistor R_1 to ground and the other through resistor R_2 and capacitor C_4 to ground. The capacitances of C_3 and C_4 are so chosen that they will offer a low reactance at the i.f. compared with the resistance of R_1 and R_2. This means that practically all of E_1 also appears across R_1 and R_2. So far as the high-frequency i.f. is concerned, R_1 and R_2 are both effectively in parallel with L_1, and the reference voltage E_1 also appears across the load resistors R_1 and R_2.

Since R_1 is the load resistor for D_1, both voltages E_1 and E_2 act on diode D_1; and, since R_2 is the load resistor for D_2, both E_1 and E_3 act on diode D_2. The on- and off-resonance conditions function as in the conventional discriminator. The audio output appears across C_4 or between the high side of R_4 and ground. In this circuit, resistor R_3 and capacitor C_6 represent a de-emphasis network, and the de-emphasized audio is taken across C_6. The triode section of the tube is used as the first audio voltage amplifier.

In the discriminator circuit of the Consolidated Television model 2315, a resistor is used in place of the usual coupling coil L. This modified circuit is very similar to that of the Du Mont model RA-102 discussed previously. The interesting and unique thing about this network is the method of detection used. For the detection process two diodes are still employed, but they are two germanium crystal diodes instead of the conventional duo-diode tube.

The circuit for this arrangement appears in Fig. 5-24. The two 1N34 diodes are connected in a similar fashion to the usual duo-diode tubes. Although 1N34 crystal diodes are more expensive than any duo-diode tube such as the 6H6 or 6AL5, they offer a number of advantages as compared with the tubes. A 1N34 crystal diode occupies much less space than the conventional duo-diode tube. With these crystals, sockets are not necessary and they do not require any heater voltage.

Aside from these physical characteristics, crystal diodes such as the 1N34 offer greater electrical stability and thus a better balance is obtained in the discriminator circuit. The total shunt capacitances such as the input and output tube capacitances and stray capacitances are less when using germanium crystal diodes than with conventional duo-diode tubes. This is another feature which helps toward better discriminator balance. The positive or plus terminal of the 1N34 crystal is usually the plate of the diode and the negative or minus terminal the cathode.

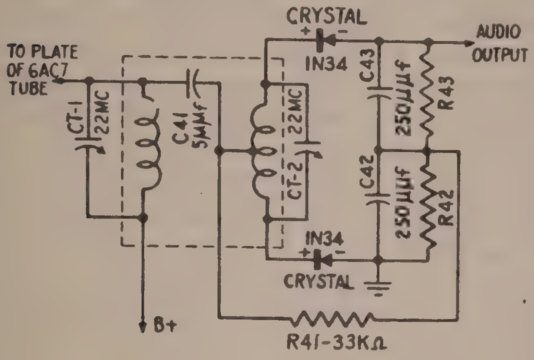


Fig. 5-24.—The discriminator circuit used in the Consolidated Television model 2315. Note that a resistor is used in place of the usual coupling coil.

Courtesy Consolidated Television Corp.

THE RATIO DETECTOR

In the previous discussion on the discriminator detector circuits, it was indicated that a limiter was necessary to produce an input f-m signal to the discriminator that was constant in amplitude. The limiter is needed because the discriminator responds to amplitude changes in the signal as well as to frequency changes. To dispense with a separate limiter tube, *the ratio detector,* was developed. This detector responds to frequency changes only and not to amplitude changes in the input system. Although its operation is different from that of the discriminator detector, it is nevertheless somewhat similar in circuit analysis. This will be seen as we progress with the analysis of the detector.

Simplified Ratio Detector

To understand fully the operation of a typical ratio detector circuit let us first study a simplified version of such a circuit as illustrated in Fig. 5-25. From this diagram we notice one thing that is common to all ratio detector circuits, namely, that the two diodes used are wired in *series aiding* with respect to the load instead of *series opposing* as in the discriminator detector circuit. By tracing the circuit it will be found that the plate of one diode is connected to the cathode of the other diode through the secondary of the transformer. The plate of the latter in turn is connected to the cathode of the former diode through a battery. Compare these connections with that of the discriminator detector in Fig. 5-8, and the difference will be immediately apparent. This difference in circuit arrangement of these two detectors is one quick method of telling them apart.

Coming back to Fig. 5-25, we find that the transformer network in conjunction with C and L is similar to the discriminator detector arrangement in that the voltage across the primary, L_1, is also across L. This voltage in conjunction with the individual voltages across L_2 and L_3 is effectively on each respective diode. Across the output of this circuit appear two capacitors C_2 and C_3 and in parallel with them a battery of fixed voltage, E_B. In the discriminator detector, besides two capacitors, two resistors also existed, across which the differential output voltage was developed. However, when any amplitude changes occurred in the input signal to the discriminator detector, the output voltages across each resistor changed, making the differential output voltage different, indicating that the detector was responsive to a-m signals as well as to f.m.

In the simplified ratio detector circuit of Fig. 5-25, the voltage from diode to diode in the output side of the circuit is maintained constant at voltage E_B by the battery. Consequently, the total voltage

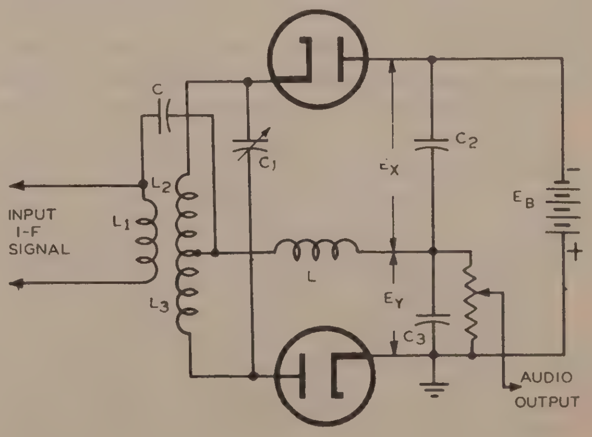


Fig. 5-25.—Simplified schematic of a ratio detector circuit. Note that the diodes are connected in *series aiding* rather than series opposing as in the discriminator detector.

across capacitors C_2 and C_3 will always be equal to E_B. According to the polarity of the battery connection, no current will flow in the circuit until a signal is applied. The d-c path of this output circuit is through the battery, and the a-c path through the two capacitors C_2 and C_3. When an f-m signal is detected by this arrangement, the individual voltages E_X and E_Y across capacitors C_2 and C_3 respectively, will be constantly changing due to the change of deviation of the f-m signal, *but* their sum will be constant due to the battery voltage E_B. At all times $E_B = E_X + E_Y$. Since the individual values but not the sum of E_X and E_Y can change, it is their *ratio* which will be constantly changing, and by placing a potentiometer across C_3 the audio modulation of the f-m signal can be tapped off this resistance. This will all be clearer when a typical ratio detector circuit is analyzed.

Any amplitude variations in the input signal will not appear across either C_2 or C_3 as changes in voltage due to the constant voltage across the output due to the battery. Only *frequency* changes will appear across both capacitors. This is different from the discriminator detector circuit, where both amplitude and frequency changes were recorded as voltage changes across the load *resistors*. This simplified form of the ratio detector was analyzed first because through the use of the battery we were easily able to show the fixed voltage across the output and why this detector responds only to frequency variations.

Practical Ratio Detector

The use of a battery for constant voltage output in the ratio detector for f-m receivers is not practical because, due to the nature of the incoming f-m signals, we desire to have the output voltage constant only at the average strength of the incoming signal. Since the sound carriers of the different television stations are not all of the same strength and also since the effective strengths of the individual carriers at the receiver change, it was found that the best thing possible was to have a relatively constant voltage across the output determined by the average value of the incoming signals. This was accomplished by placing a parallel *RC* network in the output of the circuit instead of the battery, as shown in the typical ratio detector circuit of Fig. 5-26. Let us now study this circuit in conjunction with the new *RC* output circuit, and the over-all action of this detector will become apparent.

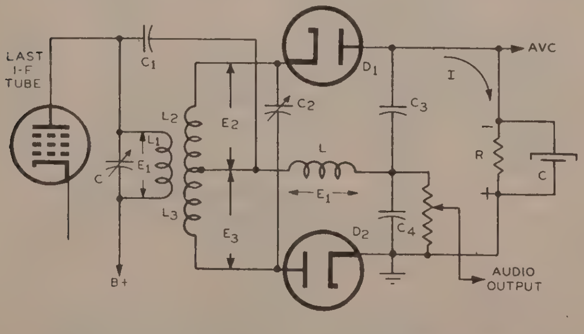


Fig. 5-26.—A typical ratio detector circuit. A parallel RC network has replaced the battery of Fig. 5-25 in the output of the circuit.

This new circuit is very similar to the discriminator detector arrangement in many ways. For instance, the voltage E_1 can be considered as effectively the same, even though it is not obtained from the limiter. In both circuits the signal is of the i.f. This voltage E_1 is coupled to the rest of the circuit in two ways: by induction into L_2 and L_3 and by capacitive coupling across C_1 to L and C_4. The reactances of C_1 and C_4 at the i.f. are negligible compared with the reactance of the choke L. Consequently, the voltage drop across L is likewise E_1 as seen in Fig. 5-26. Depending on the degree of coupling between the primary tuned circuit and the secondary tuned circuit, a certain amount of voltage is coupled across each half of the center-tapped secondary. So far as voltages E_2, E_3, and voltage E_1 across L are concerned, they all function in *exactly* the same manner as the corresponding voltages in the discriminator detector circuit of Fig. 5-8. This is true, too, for the vector diagrams of Figs. 5-9 and 5-10 and Figs. 5-12 through 5-14 with respect to the applied voltages to the individual diodes during the constant frequency changing of the incoming f-m signal. Thus, the same phase shifting process is employed in the ratio detector circuit.

Examining Fig. 5-26 once more, it is evident that since the diodes are connected in series aiding they draw current in the same direction relative to R, which is also in series with them. Consequently, using the convention for the flow of electrons from cathode to plate, the current I will follow the path indicated by the arrow, and the top part of resistor R will become negative with respect to its bottom or grounded end. If the primary and secondary tuned circuits are both resonant to the i.f. and if an unmodulated i-f carrier signal (of the same frequency) is injected into the circuit from the i-f amplifier, the two capacitors C_3 and C_4 both will be charged to the same voltage due to the symmetry of the circuit.

Now if the i-f carrier were frequency modulated, the voltages appearing across capacitors C_3 and C_4

would vary according to the modulation of the i-f carrier. This happens as follows: It was mentioned that rectified current would flow in the direction shown in Fig. 5-26 and that the top portion of the resistor *R* would have a negative potential on it. The values of the resistor *R* and capacitor *C* are so chosen that they represent a long time-constant network. Usually the value of this time-constant network can vary anywhere between one-tenth and one-quarter second and still be effective to the desired degree. Consequently, with a long time constant, it will take the capacitor *C* quite some time to discharge fully through *R*. Therefore, the negative voltage at the top of resistor *R* will remain practically constant over the range of the lowest audio frequency desired to be reproduced in the output of the set. In other words, a time constant of one-tenth of a second corresponds to the period of a frequency of 10 cycles per second; therefore, for frequencies above 10 cycles per second, the duration of one cycle would be shorter than the time constant, and so the voltage across the *RC* combination will remain practically constant. (The higher the audio frequency, the shorter the duration of one cycle.)

Since the voltage across *R* and *C* is constant, the sum of the voltages across C_3 and C_4 must remain constant. However, if the carrier frequency falls below, or rises above, the i.f., the voltages appearing across C_3 and C_4 will differ in value according to the degree of off-resonance condition of the i-f signal. No matter what the difference between these voltages, their sum always remains the same, but their ratio will be varying at the rate of the deviation of the f-m signal, and it is this change in ratio which is detected. If the i-f signal is frequency modulated, the i.f. will vary above and below its resonant frequency according to the degree of f.m. This accordingly will vary the voltages appearing across C_3 and C_4, but in a certain proportion determined by the potential across the *RC* combination. Consequently, it can be said that the voltage across C_4 varies at an audio rate (due to the degree of f.m.) Therefore, the a-f output may convenientlv be taken off across C_4, because one side is grounded. and applied to the audio section of the set. The instant when the incoming f-m i-f signal is at the exact resonant frequency of the tuned circuits of Fig. 5-26, the a-f voltage across C_4 will be zero. At instantaneous frequencies of the incoming f-m i-f signal above and below the resonant frequency of the tuned circuits, the voltage across C_4 will vary at a rate determined by the changing frequency of the f-m signal. Since the f-m signal is changing, or is being deviated, at an audio rate (that is, at the rate of its modulating signal), the output voltage across C_4 will be varying at an audio rate. In this type of circuit the voltage appearing across C_3 will be larger than that across C_4 at frequencies below the i.f., and above the i.f., the voltage across C_4 will be larger than that across C_3.

The basic part of the ratio detector that removes any a.m. appearing in the input is the *RC* time constant network of Fig. 5-26. It is the constant voltage across resistor *R* and capacitor *C* that plays the primary role in the removal of a.m. Let us suppose that an a-m signal appears at the input of the ratio detector and see what happens:

Any a-m signal will tend to increase the voltages across capacitors C_3 and C_4. However, the voltage across the *RC* network cannot change rapidly enough to follow the a.m., due to the nature of the long time constant, and the a.m. therefore, cannot change the voltage across C_3 and C_4. In other words, the capacitor *C* charges or discharges so slowly through *R* that the potential at the top of resistor *R* (or the plate of diode D_1) remains nearly constant, and any a.m. cannot change the voltage across capacitor *C* in step with this a.m. Consequently, sudden increases in amplitude of the f-m carrier will have no effect in the output audio circuit, because these sudden increases of amplitude cannot appear across either C_3 or C_4 as a change in voltage.

AVC From Ratio Detectors

In the limiter discriminator arrangement, avc voltage was available in the grid circuit of the limiter due to grid rectification action. In the ratio detector system, since limiters are not employed, the avc is obtained from some other place. In the ratio detector circuit of Fig. 5-26 it is noticed that the voltage across resistor *R* serves as a means of obtaining avc voltage. Since the time constant network of *RC* is made to produce a constant output voltage at the average strength of the incoming signal, it is readily evident that this output voltage will change in accordance with the varying average strength of the incoming signal.

What this means is that the capacitor *C* in conjunction with resistor *R* averages these signal strength changes appearing across *R*. The time constant is not considered large in this instance as compared with the length of time required for changes in average signal strength. However, the effect of the time constant is sufficiently large to produce effective removal of sudden changes in a.m., including that brought about by the response characteristics of the i-f stages. This is possible because the input signal does not change in strength as rapidly as these other amplitude variations, and the *RC* combination permits *slow* changes in voltage in accordance with slow changes in the received signal. Therefore, the negative voltage at the top part of resistor *R* serves as a source of avc voltage.

Ratio Detector Modifications

The circuit of Fig. 5-26 was one of the first commercial types of ratio detector circuits to be used. However, most of the ratio detector circuits that are used in the sound channel of today's television receivers are modifications of the one just discussed. The circuits of Figs. 5-25 and 5-26 were both analyzed in order to illustrate the similarity to the discriminator circuit.

The most difficult part in the design of the ratio detector circuit is the transformer, namely that comprising coils L_1, L_2, and L_3 in Fig. 5-26. It is beyond the scope of this book to go into such design work, but it should be remembered that such factors as the proper coupling between windings, the respective Q's of the coils both during the diode unloaded and loaded conditions, and the gain of the last i-f stage are important in this transformer design.

In the following are discussed a number of different types of ratio detector circuits. The modified ratio detector circuit that appears in Fig. 5-27 is from the Andrea chassis VJ12. It may not be readily evident from this drawing how the signal voltage across the transformer primary appears across each diode.

Unless something is known about the construction of the transformer comprising *L*, L_1, L_2, and L_3, it is somewhat difficult to understand this circuit. It should be known first that, due to inductive coupling, voltages E_2 and E_3 appear across L_2 and L_3, respectively, as in other types of detectors. The important thing that should be noted is that coil *L*, which has only a few turns, is a separate winding usually closely wound around or near the bottom or B plus side of coil L_1. In this manner the coupling between these two coils is a maximum, and practically all of voltage E_1 appears across this coil and the series capacitor C_5 to ground. Since *L* is untuned, the voltage induced into it from L_1 is 180° out of phase with E_1. The voltages E_2 and E_3, as just pointed out are in quadrature with E_1. Therefore, these two voltages are also in quadrature with the voltage across L, as required.

Examining the circuit of Fig. 5-27 a little more closely it will be seen that one end of coil *L* is tied to the junction of the two 35 $\mu\mu f$ capacitors, *C*111 and *C*112, and the other end to the center tap of the transformer secondary. With this in mind we can readily see that coil *L* is common to both diodes. Consequently voltage E_2 and that across *L* are applied across diode D_1 and voltage E_3 and that across *L* are applied across diode D_2. The vector diagrams illustrating the operation of this circuit at resonance and off-resonance conditions are very much the same as those discussed with the discriminator circuit. As far as the voltage vectors acting on the two diodes are concerned, the voltage vec-

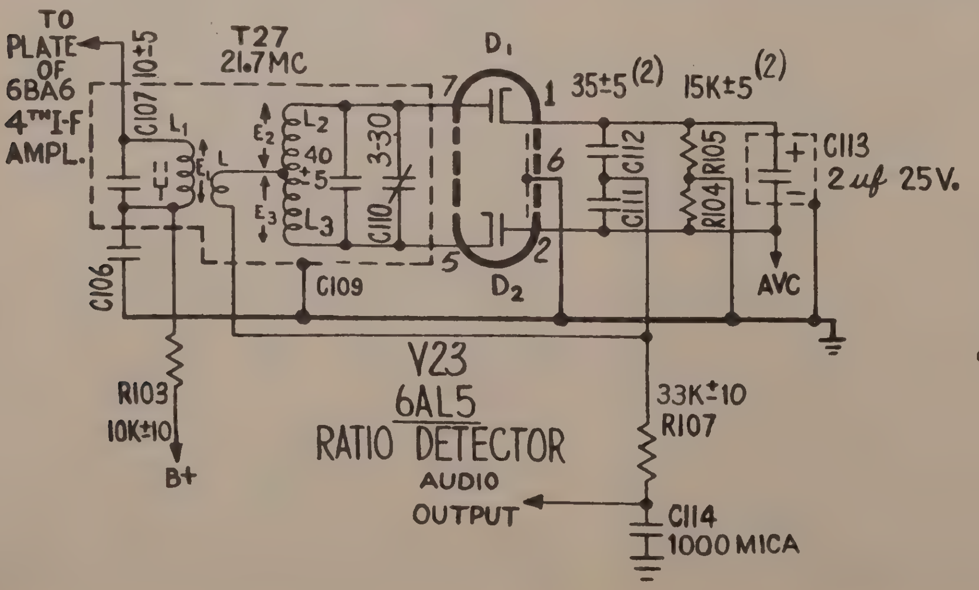


Fig. 5-27.—Modified ratio detector circuit as used in the Andrea television chassis VJ12.

Courtesy Andrea Radio Corp.

tor representing that voltage drop cross L is 180° out-of-phase from that normally indicated in the previously discussed detector circuits. However, nothing is radically changed because all this means is reversing the two horizontal axes of the previous vector diagrams and in this process the phase relationships between the voltages and current remain the same.

We know from previous ratio detector analysis that at the center frequency of the i-f signal, the voltages across the balanced output capacitors are equal. We also know that for frequencies different from the center i.f. (i.e. off-resonance conditions) the voltages across the capacitors will no longer be equal but will vary in accordance with the varying frequency of the f-m i-f signal. Consequently, the junction point between these two capacitors varies in potential with respect to the off-resonance conditions and is a convenient place to obtain the audio output signal.

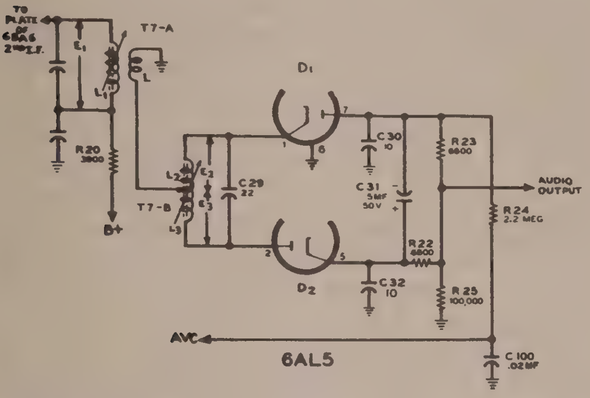


Fig. 5-28.—The modified ratio detector circuit appearing in the Belmont model 22A21.

Courtesy Belmont Radio Corp.

Looking at Fig. 5-27 once again, we see that the two balanced output capacitors are $C111$ and $C112$. Across these two capacitors is the output load resistance composed of two separate but equal resistors, namely $R104$ and $R105$, with their junction point grounded. This means that when the incoming f-m signal is equal to the resonant frequency of the transformer, the junction of capacitors $C111$ and $C112$ is effectively at ground potential. By connecting a suitable coupling network from the junction of $C111$ and $C112$ we have a means of obtaining the audio output signal. In Fig. 5-27 this coupling network is seen to consist of resistor $R107$ and capacitor $C114$ with the audio voltage signal output taken off the capacitor.

Another modified form of the ratio detector appears in Belmont model 22A21, the detector circuit of which is shown in Fig. 5-28. This circuit is radically different because coils L, L_1, L_2, and L_3 are not encased in one shield can as is usual. Part number $T7$-A contains only the primary (L_1) and tertiary (L) coils while $T7$-B contains only the so-called secondary. Actually there is little or no inductive coupling between these two units because they are separated at quite some distance on the chassis. Therefore the voltages that appear accross L_2 and L_3 are not due to transformer action as in the previously discussed circuit. Coil L_1 is closely coupled to coil L and by transformer action voltage E_1 appears across L but of opposite phase, similar to the circuit of Fig. 5-27. From the diagram of Fig. 5-28 it may not be readily evident how voltages E_2 and E_3 appear across their respective inductances L_2 and L_3. In order to see this more clearly that part of Fig. 5-28 to the right of and including coil L has been redrawn in Fig. 5-29.

We know that when a signal is being received, a voltage will be induced into coil L from coil L_1 which is approximately of the same magnitude as that across L_1 because of the close coupling. Since coil L has one end tied to the center connection of coils L_2 and L_3 and the other end grounded, the voltage appearing across it also effectively appears across each diode. The completed circuit is through the ground connection between the balanced capacitors $C30$ and $C32$.

Let us analyze the action of the circuit on only one half cycle of the input signal at a time. When the voltage across L is on the positive half cycle, the plate of diode D_2 and the cathode of diode D_1 will both have a positive potential and the cathode of diode D_2 and plate of diode D_1 will both have a respective negative potential. Thus diode D_1 will not conduct because its plate *is less* positive than its cathode. However, diode D_2 will conduct because its plate *is more* positive than its cathode. Consequently a current will flow in coil L_3 and a voltage drop E_3 will appear across this coil. This voltage drop is due to conduction current flowing through it and not to induction as was usually the case. How-

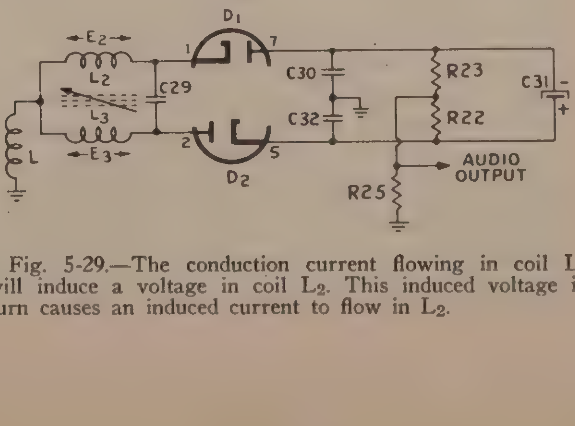


Fig. 5-29.—The conduction current flowing in coil L_3 will induce a voltage in coil L_2. This induced voltage in turn causes an induced current to flow in L_2.

ever, coil L_2 and L_3 are usually so closely wound together on one form (it usually is single center-tapped coil) that there is effectively unity coupling between them. The *conduction current* flowing *through* L_3 will thus set up magnetic lines of force which will link coil L_2 and induce a voltage in coil L_2 of the same magnitude as that across L_3. Thus voltages E_2 and E_3 equal each other in magnitude. However, due to the method of connection between L_2 and L_3, the two equal voltages E_2 and E_3 are 180° out-of-phase with each other. Therefore we have the same phase conditions between the three voltages, (that across L and E_2 and E_3) as in previous circuits.

The induced voltage across E_2 will then cause an *induced current* to flow through its circuit. Consequently on the positive half cycle we see that equal but 180° out-of-phase voltages appear across secondary coil L_2 and L_3. On the negative half cycle of the signal across L, the potentials across each diode are interchanged from what they were on the positive half cycle. This means that diode D_1 will conduct first and coil L_2 will have *conductive current* flowing through it and thus coil L_3 will have an *inductive current* flowing through it. The voltages across L_2 and L_3 are the same as on the positive half cycle because each half cycle is of the same amplitude. It is thus seen that during each cycle of input signal equal but 180° out-of-phase voltages appear across L_2 and L_3.

The phase relationships between the voltage across L and that across L_2 and L_3 are the same as in the circuits previously discussed.

Slope Detection

In the first commercial television receivers the sound channel employed a.m. However, when the FCC changed the sound television transmission from a.m. to f.m, the sound channel of the television receivers likewise had to be changed in order to receive f-m sound signals. The only commercially used f-m detector network at the time of the changeover was the limiter-discriminator circuit. To convert the a-m sound channel to one capable of detecting f.m. would require many changes, especially in the detector circuit. In order to avoid rewiring and adding and deleting components, the circuits already in the sound channel were uti'ized but were detuned so that detection of an f-m signal was possible.

In other words, the only essential change from the a-m circuit to one of so-called f.m. was a process of detuning the i-f transformers in the set. This process of detuning was very important for the correct operation of the sound channel on f-m signals. This process of detuning was done to such television receivers as the RCA models TRK-9 and TRK-12 and the General Electric models HM-171 and HM-185.

Detuning of the i-f transformers changed the response curve of the circuits in question. The detuning was such that either one or both of the sides of the response curve became quite linear. The more linear the sloping characteristic of the response curve, the less the amount of distortion in the output. The amount of detuning was such that the center frequency of the incoming f-m i-f signal fell on the linear portion of the sloping characteristic of the response curve. (Either the slope at the high or low frequency end of the response curve can be used. Of course the more linear the slope, the better the detection). The incoming f-m i-f signal will be changing in frequency an equal amount on either side of its center frequency, and the i-f transformers are detuned to the point where the frequency swing of this signal will fall upon the linear slope of the response curve. By the f-m i-f signal falling on the sloping characteristic of the i-f response curve, detection of the f-m signal is possible. This type of detection is commonly known as slope detection.

In order to show how an f-m wave can be detected by this method, let us refer to the response curve of Fig. 5-30. This curve approximates the selectivity curve of an i-f transformer with somewhat broadband characteristics. In this drawing we assume that the frequency represented by point A on the left slope of the curve and the center frequency of the incoming f-m i-f signal are the same. The curve on the bottom of the drawing is just a sine wave depicting the frequency variations of the input f-m signal. This curve *is not* an f-m wave. Since we know that an f-m wave varies in frequency we can illustrate this frequency variation by a sine wave that is contant in amplitude. Thus the peak amplitude of the sine wave at the bottom of the drawing illustrates the peak deviation of the incoming f-m i-f signal.

From this drawing it is readily noticed that the peak-to-peak frequency deviation of the f-m input signal is well within the so-called linear range of the left hand portion of the curve. Since the curve is sloping, conversion of the frequency variations of the f-m signal into one that also varies in ampli-

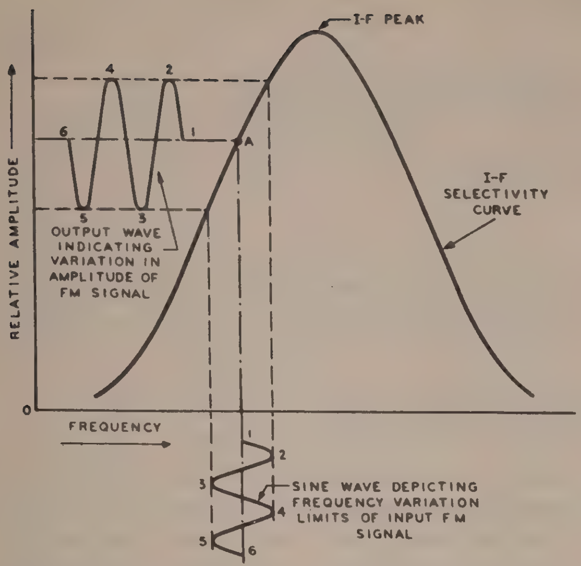

I-F PEAK
4
2
6
1
A
5
3
I-F SELECTIVITY CURVE
OUTPUT WAVE INDICATING VARIATION IN AMPLITUDE OF FM SIGNAL
RELATIVE AMPLITUDE
0
FREQUENCY
1
2
3
SINE WAVE DEPICTING FREQUENCY VARIATION LIMITS OF INPUT FM SIGNAL
4
5
6

Fig. 5-30.—When an f-m i-f signal falls on the sloping characteristic of the i-f response curve, detection of an f-m signal is possible.

tude is brought about. The frequency variation of the f-m signal changes in accordance with the amplitude of the audio modulation. For the sake of simplicity we showed the frequency variation as constant, but for speech and music it is anything but constant. The output wave at the left hand side of the drawing illustrates the amplitude change of the f-m signal brought about by the slope of the response curve. It is shown as being constant in amplitude because the incoming f-m signal is shown as being constant in frequency deviation over a short period of time.

The output signal from the i-f transformer network will still be an f-m signal but this f-m signal will also be varying in amplitude as well as in frequency. The conversion of the input signal is such that the output f-m signal varies in amplitude in accordance with the ratio and strength of the audio modulating signal.

The usual diode detector following the i-f stages of a-m systems will not detect true f-m waves. However, with the detuning method just described, the input signal to the diode detector will vary in amplitude as well as frequency. Consequently the diode detector will respond to the amplitude variations and detection will be complete. The audio signal output from the detector will thus be varying at the frequency of the audio modulation of the f-m signal.

CHAPTER 6

THE VIDEO I-F AND DETECTOR SECTION

By Seymour D. Uslan

In the regular a-m and f-m receivers, the i-f stages provide the set with the necessary selectivity and gain for proper reception. The selectivity of the i-f stages is primarily determined by the i-f transformers and the gain of these stages by the i-f amplifiers. In a-m broadcasting, the bandwidth is very narrow compared to that for f.m. In either case the tuned i-f circuits of the receivers are designed to pass only those frequencies within the bandwidth of the desired signal. By bandwidth is meant those frequencies involved in the transmitted sidebands. In a-m broadcasting (to 1500 kc) the bandwidth of the signal can be no greater than 10 kc, or 5 kc on either side of the carrier. In f-m broadcasting the over-all bandwidth is 150 kc, or 75 kc on either side of the carrier frequency. In the f-m sound channel of television, the over-all bandwidth is 50 kc (see chapter 5). For the video part of the television signal, as previously mentioned, the bandwidth far exceeds any of the other types of signals due to the very high frequencies involved in the video signal.

The video i-f system of television receivers has to perform the same function of selectivity and gain as in the other sets. However, due to the bandwidth of the signal involved, the design of the i-f circuits greatly departs from that used in regular a-m receivers and even differs appreciably from f-m receivers. The video i-f amplifier must handle frequencies extending over 5 mc. Not only must the i-f amplifiers and associated circuits have a flat response over this range, but they must also reject interfering signals close to the edges of the pass band. The coupling circuits between the video i-f amplifiers are usually not as simple as the i-f transformers in the a-m and f-m receivers. Many different i-f circuits will be discussed in this chapter in conjunction with the methods of obtaining the required bandwidth and interfering signal rejection.

The detector stage of the video section of a television receiver follows the i-f stages. Its function is the same as that of a detector employed in a-m receivers and similar type circuits are used. The diode detector is a common form of detector circuit. Slight modifications in circuit arrangement are required due to the high frequencies involved and also due to the required polarity of the video signal on the picture tube.

The Video I-F System

The output signal from the mixer tube is a combined video and sound i-f signal. One of the functions of the i-f system is to deliver the sound modulating frequencies at the input to the audio amplifiers and the video modulating frequencies at the input to the video amplifiers. The methods of obtaining this, however, differ in many of the receivers.

For instance, some receivers tap off the sound i-f signal at the output of the mixer stage and apply it directly to the f-m i-f amplifiers; the video i-f signal is also sent through its i-f system. In other receivers the combined i-f signals at the output of the mixer are first sent through one or two i-f amplifiers which amplify both signals before they are separated for their respective channels. These tubes are sometimes called video i-f amplifiers; however, they are also sound i-f amplifiers and help increase the gain of the sound i-f signal. Still other receivers send both i-f signals through a complete series of amplifiers, but the reason in this case is not just to amplify both i-f signals but also to produce a new sound i-f. All these circuits will be discussed later on.

The terminology of the *video i-f stages* is applied to all stages that have to pass the video signal, whether or not the sound i.f. is also involved. In other words, it is common practice to refer to those i-f stages *not* used solely for the sound i.f.'s as video i-f stages. It is in these video i-f stages that we are interested. Besides providing the desired gain and broad selectivity, the over-all flat response curve of the video i-f system must have a special sloping shape around the video i-f carrier part of the curve. In general the shape of the correct over-all i-f response curve is quite critical. The necessary broad bandwidth characteristic of the video i-f stages is obtained primarily by two methods —overcoupled i-f stages or stagger-tuned i-f stages. Both types are used today. Before we discuss them we will study the bandwidth requirements.

Bandwidth Requirements

In order to reproduce all the video frequencies, the bandwidth of the video i-f circuits has to be quite wide. With vestigal-sideband transmission and with a maximum of 4-mc video frequency for one sideband, the maximum video i-f bandwidth should be about 5.25 mc. This was indicated in the over-all transmitted bandwidth picture of Fig. 2-4. The receiver bandwidth for the video modulated signal was illustrated in Fig. 2-6. Due to the use of vestigal-sideband transmission, the lower video frequencies are emphasized, and to avoid this overemphasis, the selectivity characteristic of the receiver for the video signals should have a sloping characteristic such as illustrated in Fig. 2-6. This sloping characteristic is in the neighborhood of the video carrier. The de-emphasis of the lower video signals is brought about by the video i-f system. Due to the process of frequency conversion, the video receiver response characteristic is reversed from that shown in Fig. 2-6.

The ideal over-all frequency response characteristic of the video i-f stages is illustrated in Fig. 6-1. The sharp cutoff characteristic on the left-hand side of the response is primarily due to the sound traps, which we will discuss later on. The right-hand side of this response should essentially be a mirror image of the left-hand side of the response shown in Fig. 2-6. Note the change in position of the video and sound carriers. In order to understand this change, let us analyze the system of frequency conversion.

In the transmitted television signal, the sound carrier is 4.5 mc higher than the video carrier. That is the reason that the sloping characteristic around the video carrier of Fig. 2-6 is on the low-frequency side of the curve. The oscillator in a television receiver is tuned above the incoming television signal. Hence, after frequency conversion, the center frequency components of the produced i-f signals are reversed in frequency spectrum relationship as compared to their carrier frequencies when transmitted.

For example on the first channel (44 to 50 mc), the picture carrier is 45.25 mc and the sound carrier is 4.5 mc higher at 49.75 mc. Assuming that the receiver oscillator is tuned to 71 mc, the i.f.'s produced will be 71 less 45.25 mc or 25.75 mc for the video signal, and 71 less 49.75 mc or 21.25 mc for the sound carrier of the same channel. Consequently we see that due to conversion action, the new carrier frequencies now termed the i.f.'s are such that the video i.f. is higher than the sound i.f. The frequency difference between them is still, however, equal to 4.5 mc. Thus in Fig. 6-1 we see the reason for the shape of the *over-all* i-f selectivity curve of the video i-f system.

If television receivers are to be selective for all video frequencies and nonselective to others, the shape of the curve of Fig. 6-1 should be adhered to as closely as possible. If this bandwidth is not strictly adhered to, the low video and high video frequencies will not be amplified equally. The low video frequencies contribute to the background of the picture and to the resolution of large objects. If the low frequencies are discriminated against, the background and large objects will not appear clear on the picture tube. This will be noticed no matter what the size of the pattern on the picture tube. Consequently, it is very important that the low video frequencies be reproduced according to the curve of Fig. 6-1.

If the bandwidth of the response curve of Fig. 6-1 were decreased by reducing the high video frequency end of the band, not as much trouble would be encountered as in the case of low video frequency discrimination.

The high video frequencies are not too critical because they add to the finer detail of the picture. A loss in the finer detail of the reproduced picture will not be noticed as readily as the loss due to low video frequencies. This is even more pronounced when the size of the screen used is small, because the smaller the picture, the more difficult it is to note the fine details. For small picture tubes a high-frequency bandwidth characteristic as wide as shown in Fig. 6-1 is not absolutely necessary for good picture reproduction. There are television receivers where the bandwidth of the video i-f system is not as wide as illustrated in Fig. 6-1, and the response at the high video frequencies is decreased. Such receivers employ small size picture tubes and the loss of the high video frequencies is not easily noticeable upon the reproduced picture.

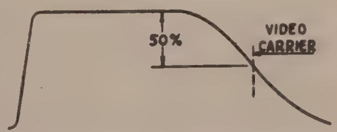


FIG. 6-1.—The ideal over-all frequency response characteristic of the video i-f stages: The sharp cutoff characteristic on the left results from the sound traps; the right-hand side of the response should essentially be a mirror image of the response shown in Fig. 2-6.

The wider the bandwidth of the i-f stages the greater will be the gain required and, consequently, the more video i-f amplifiers needed. Thus the amount of video gain, and hence the number of video amplifiers needed, depends upon the bandwidth of the i-f stages.

In the following sections we will study the different methods used in television receivers in order to obtain the broad bandwidth characteristic.

Overcoupled I-F Transformers

The use of overcoupled i-f transformers to produce the necessary wide band is quite common in many of today's television receivers. For sharply selective circuits, the response curve is usually a single-peaked curve as shown by curve *A* in Fig. 6-2, where the peak is at the resonant frequency of the circuit. Such circuits have a low coefficient of coupling. When the coefficient of coupling is increased, a point will be reached where any further increase will result in a double-peaked curve. This point is called the point of critical coupling and curve *B* in Fig. 6-2 illustrates a critically coupled curve. Note that it has a somewhat broader peak than curve *A*. When the coefficient of coupling is increased beyond its critical value, the response curve will become much broader, with double peaks appearing instead of a single peak. Curve *C* in Fig. 6-2 illustrates greater than critical coupling. The dip in the center of the curve occurs at the resonant frequency of the circuit, and the higher the coefficient of coupling, the deeper will be the dip and the greater will be the distance between the peaks. Such curves as *C* in Fig. 6-2 are desired for broad-band characteristics and are often acquired by means of overcoupling the i-f transformers.

In some systems the coupling between the coils (both tuned to the same resonant frequency) remains constant and the circuit *Q's* are varied to effectively result in a double-peaked broad-band curve. Another method of obtaining a wide band curve is to detune both the primary and secondary of a coupled circuit, with the coupling remaining constant. Overcoupling the circuit is primarily used in video i-f circuits.

It had sometimes been found that in order to obtain the necessary bandwidth, the overcoupled circuit produced a double-peaked curve that was wide enough, but the dip in the center of the curve was too low. If the transformer were loaded down and the *Q* reduced without appreciably changing the bandwidth and resonant frequency of the circuit, then a flatter response characteristic would result. In many receivers this effect is obtained by loading down the transformers with resistance.

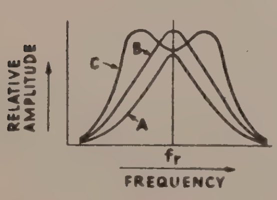


Fig. 6-2.—Response curves for different coefficients of coupling: A low coefficient produces the single-peaked curve *A*. The broader curve *B* has reached the critical point beyond which any increase in the coefficient produces a still broader but double-peaked curve *C*.

Resistance Loading

In the video i-f circuits of both the GE model 802 and Farnsworth model GV-260, which are discussed later on, loading resistors are used across the secondary of the video i-f transformers. These loading resistors reduce the over-all gain of the stage but at the same time make for a flatter and somewhat broader response.

It is known that decreasing the *Q* of a circuit flattens out the frequency response of that circuit. Placing a resistance across a tuned circuit will reduce the *Q* of that circuit. The *Q* of a circuit (or coil) is defined as the ratio of inductive reactance to series resistance. If the series resistance is increased or the inductive reactance decreased, the *Q* decreases. Placing a resistor across a parallel *LC* tuned circuit will effectively increase the series resistance in the inductive or capacitive branch. Since the coil always contains some inherent (series) resistance as compared to the negligible amount contained in the capacitor, the resistance placed across the parallel *LC* circuit is said to increase the series resistance of the inductive branch. The *lower* the value of the parallel resistance, the *greater* will be the effective series resistance and hence the *lower* the *Q*.

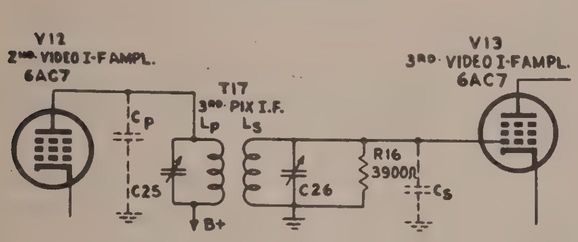
The parallel resistance, however, cannot be too low because the gain per stage decreases with a decrease in *Q* and a certain amount of gain is required for proper operation of the video i-f system. The resonant frequency of a parallel *LC* circuit also depends upon the series resistance in the circuit. By lowering the *Q* of such a circuit by the insertion of a parallel resistor, the resonant frequency is also changed. This change, however, depends upon the magnitude of the parallel resistance and the value of inductance in the circuit. If the effective series resistance becomes appreciable, the resonant frequency will change.

In the design of circuits with resistance loading, the manufacturer usually takes this into account, so that with parallel resistance loading, the desired resonant frequency is still attained. Whenever resistance loading is used to broaden out the response characteristic of a tuned circuit where the resonant frequency is high, it is always advisable when the resistor needs to be replaced that an exact duplicate be substituted; special emphasis should be placed on trying to keep the leads and wiring similar to what they were before.

Typical Overcoupled Circuits Employing Resistance Loading

We will now investigate two typical overcoupled video i-f circuits that employ resistance loading across the secondary of their i-f transformers to flatten the response characteristic further. We will not show the complete video i-f circuits of each set but only one coupling network from each, because the other video i-f stages are pretty much the same.

Fig. 6-3 illustrates the video i-f transformer coupling network used between the second and third video i-f amplifiers of the GE model 802. This circuit has been drawn in a simplified form for the sake of the discussion. The primary resonant circuit consists of inductance L_P and the parallel combination of capacitances C_P and C_{25}. The secondary resonant circuit consists of inductance L_S, the parallel combination of capacitors C_S and C_{26}, and resistor R_{16}. Capacitors C_{25} and C_{26} are a physical part of T-17, the third video i-f transformer, and are variable. (The video i-f transformers are often called *pix* i-f transformers as in this model under discussion. In fact, the word *pix* is used quite often in place of the word video.) Capacitance C_P represents the total shunting capacitances common to the primary circuit of the transformer and essentially consists of the output capacitance of the second video amplifier, the distributed capacitance of coil L_P, and the stray wiring capacitances in the vicinity of the primary circuit. In similar fashion, capacitance C_S represents the total shunting capacitances common to the secondary circuit of the transformer and essentially consists of the input capacitance of the third video i-f amplifier, the distributed capacitance of coil L_S, and the stray wiring capacitances in the vicinity of the secondary circuit.



·Courtesy GE

FIG. 6-3.—Simplified circuit diagram of the video i-f transformer network used between the second and third video i-f amplifiers of the GE model 802.

Coil L_P is closely coupled to coil L_S so that the transformer circuit is effectively overcoupled and produces a broad response. The response is made still broader by the use of a 3900-ohm loading resistor R_{16} across the secondary of the transformer. This reduces the Q of the secondary circuit and hence flattens and broadens the response characteristic. Normally when a circuit is overcoupled, two peaks of the same amplitude will appear equally spaced on either side of the resonant frequency of the response curve. If the Q's of both the primary and secondary resonant transformer circuits are reduced by the same amount, the double-peaked response characteristic will be flattened out evenly.

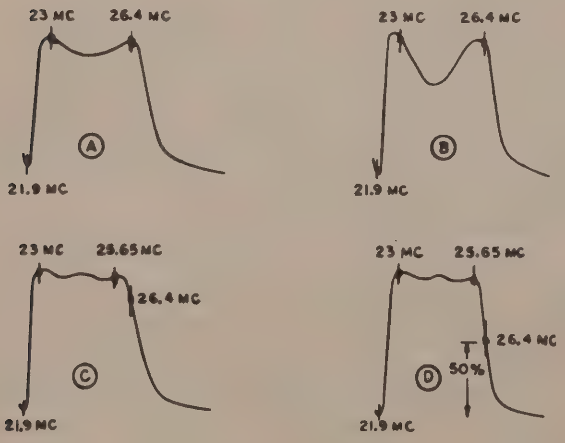
However, if the Q of only one circuit (the secondary in the circuit of Fig. 6-3) is reduced, the peaks of the overcoupled curve will not be reduced evenly. The degree of overcoupling, the amount of reduction in Q, and the change in the impedance that the secondary reflects into the primary due to the reduction in secondary Q, determines how much each peak will be reduced. The change in Q of the secondary circuit due to the resistance loading injects a different reflected impedance into the primary circuit. This new reflected impedance changes the Q of the primary circuit, but this change of Q is not the same as that of the secondary circuit.

For most practical purposes, those television receivers using overcoupled video i-f transformers with resistance damped secondaries are usually designed so that there is very little change between the damping of the two peaks of the overcoupled response curve. Not all the overcoupled video i-f stages in any one receiver produce exactly the same response curve. They have to vary slightly in order to help produce an over-all video i-f response that resembles curve *C* of Fig. 6-2. Each video i-f stage contributes to the gain and frequency response to produce the necessary response curve for proper operation of the receiver. It should be remembered that resistance loading reduces the gain of a stage, and the video amplifiers must increase the gain to the necessary amount for correct operation of the receiver.

If all the video i-f circuits produced the same overcoupled response curve, the over-all i-f curve would be of an undesirable nature. The curves are additive, resulting in a greater difference between the amplitudes at the peaks and the trough; thus the over-all curve will be less flat than any one of the original curves. In order to avoid this, four video i-f stages are employed in the GE model 802 where three are overcoupled stages each resulting in a double-peaked curve and one is a single-peaked stage. The first, third, and fourth video i-f transformers are overcoupled, double-tuned circuits with resistance loading in their secondaries. The second i-f stage uses a single parallel

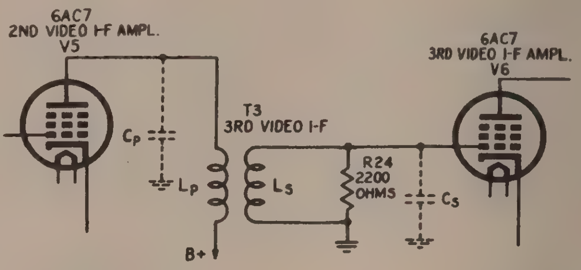
LC resonant circuit, effectively contributing a response curve that has only one peak.

The curves of Fig. 6-4 will make this somewhat clearer. These curves represent the appearance of the video i-f signal of the GE model 802 at different points in the i-f system. Curve *A* represents only that contributed by the last video i-f stage which is overcoupled and resistance damped, resulting in the double-peaked curve. Curve *B* is the result of the additive responses of both the third and fourth (last) video i-f stages. The third i-f transformer is likewise overcoupled and, together with the fourth i-f stage, helps produce a resultant curve that has a deeper trough than the previous curve *A*. This means that curve *B* will not allow those video frequencies that fall within the center of the curve to be amplified as well as the frequencies on either side of the trough. To obtain equal amplification, the response curve should be flat. This is accomplished by the use of the third i-f stage, which is a single-tuned stage, essentially producing a single peak in the region of the frequency where the trough of curve *B* appears. When all the response curves of the second, third, and fourth video i-f stages are added together, they result in a response that has a fairly flat top characteristic as indicated by curve *C* in Fig. 6-4. The response of the second video i-f stage, when added to the response of the third and fourth video i-f stages (curve *B*), increases the gain at the mid-frequency region, thereby resulting in curve *C*. The first video i-f stage is also overcoupled and resistance damped, but the amount of overcoupling and damping is so designed that it chiefly contributes to shaping the final resultant curve for the correct frequency response. Curve *D* of Fig. 6-4 is the over-all video i-f response curve of the receiver under discussion. The frequency markers on the curves indicate how the relative shape progresses from one video i-f stage to the other. The shaping of the left side of the curves is due to the 21.9-mc sound i-f traps of the receiver.



Courtesy GE

FIG. 6-4.—Curves representing the video i-f signal of the GE model 802 at different points in the i-f system. The double-peaked curves *A* and *B* represent, respectively, the response of the fourth i-f stage and the additive signal of the third and fourth stages. The single-peaked curve *C* represents the addition of the second, third, and fourth stages, and *D* the over-all video i-f response curve of the receiver.



Courtesy Farnsworth Tel. and Radio Corp.

FIG. 6-5.—A simplified diagram of the third video i-f stage of the Farnsworth model GV-260.

In the video i-f section of the Farnsworth model GV-260 (first production), overcoupled i-f transformers are used with resistance loading across the secondaries. A simplified drawing of the third video i-f stage of this receiver appears in Fig. 6-5. It will be noticed that in this circuit there is no separate capacitor and that the transformer is *not tunable.* The circuit is so designed that coil L_P is parallel resonant with capacitance C_P which represents the total shunting capacitances common to the transformer primary. Likewise coil L_S is in parallel resonance with C_S the total shunting capacitances common to the transformer secondary. Coils L_P and L_S are overcoupled so that a double-peaked response curve results and the 2200-ohm resistance R_{24} across the secondary reduces the *Q* of the circuit and broadens the response.

As in the GE model previously discussed, the total shunting capacitances represented by C_P and C_S include the stray wiring capacitance and the distributed capacitances of the coils, in addition to the tube capacitances. Since this transformer is untunable, these shunting capacitances are very important in determining the resonant frequency of the transformer circuit. When leaving the factory, the resonant frequency of these transformer circuits is already set. This means that anything that tends to change the values of C_P or C_S will tend to change the resonant frequency of the transformer primary and/or secondary.

Consequently if this section of the receiver is ever serviced, the person performing the servicing should be careful to replace the exact component parts that are removed (if any) and also to make sure that the wiring is in the same position as before servicing. In replacing parts, not only must they be of the same value and of the same size and make, if possible, but the length of the connecting leads and the position of the part should be the same as before. These precautions are necessary because only a minute change in the shunting capacitance will change the resonant frequency of the circuit.

If the video i-f amplifier tubes are changed, there is the likely possibility that the input and/or output capacitances of the new tube will be different from the one removed, in which case the over-all shunting capacitance will change and hence the resonant frequency of the circuit will be different. The manufacturer states that the absence of tuning adjustments in the third video i-f transformer (Fig. 6-5) and also in the second i-f transformer makes the gain of the stage higher. It also makes the over-all video i-f alignment simpler. Although all the video i-f stages in this receiver are overcoupled, the amount of overcoupling and resistance loading in each circuit is so designed that the required bandwidth is obtained without discrimination to any video frequencies. The over-all video i-f response curve of these overcoupled stages is illustrated in Fig. 6-6.

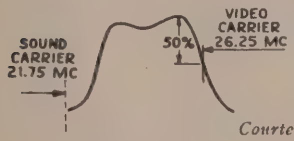


FIG. 6-6.—The over-all video i-f response curve of the overcoupled i-f transformer stages of the Farnsworth model GV-260.

Courtesy Farnsworth Tel. and Radio Corp.

Although not shown in the video i-f circuits illustrated in this section, sound i-f traps are also incorporated in these circuits to prevent interference from the sound i-f signals. These traps also help shape the over-all video i-f response curve.

Stagger-Tuned I-F Transformers

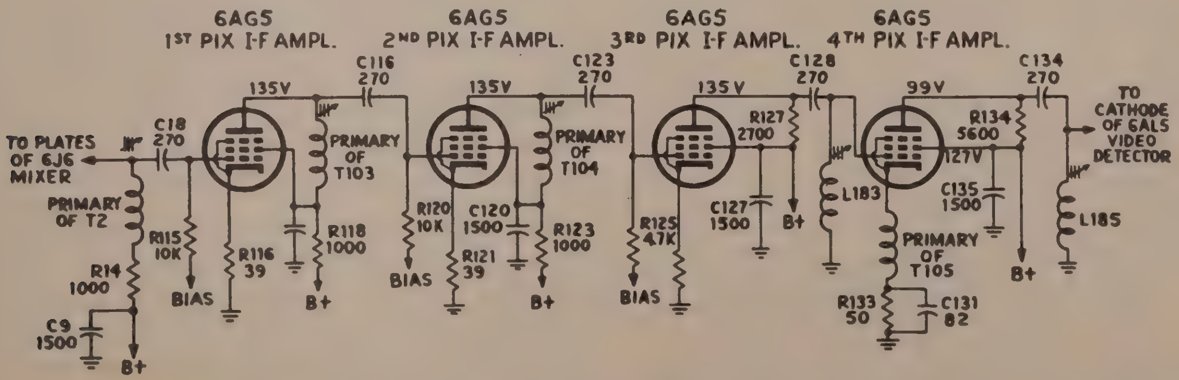
The other method of producing the necessary video i-f response makes use of *stagger-tuned* video i-f stages. In stagger tuning, the cascaded i-f stages are each tuned to a different resonant frequency. In the stagger tuning employed in the video i-f stages of today's television receivers, each i-f stage produces a single-peaked response curve, so that when all the individual response curves are combined, the resultant video i-f curve will be that required for the proper operation of the receiver.

The over-all video i-f response that results from stagger-tuned systems should approach the curve of Fig. 6-1 as closely as possible for faithful reproduction of all the video frequencies. It is possible to produce a resultant curve similar to that of Fig. 6-1 by adding together a number of curves, each single-peaked at a different frequency.

There are various methods by which the video i-f stages may be tuned. It is beyond the scope of this book to discuss every method in use today. Consequently the circuits analyzed in this section are by no means the only available types but are chosen as typical examples.

In the RCA model 630TS four video i-f amplifiers and five i-f coupling stages are employed. The coupling between stages is not by transformer action but rather by simple inductive-capacitive coupling. Fig. 6-7 is a simplified schematic of the video i-f section of the RCA model 630TS. Only those parts of the circuit and the connections of interest now have been drawn, with the traps being omitted.

In this circuit, the coils (with the exception of the primary of *T*105) are all slug tuned, so that they resonate with the distributed capacitance common to each coil and the 270 $\mu\mu f$ coupling capacitors (*C*18, *C*116, *C*123, *C*128, or *C*134). The video i-f signal voltage de-


6AG5
1ST PIX I-F AMPL.
6AG5
2ND PIX I-F AMPL.
6AG5
3RD PIX I-F AMPL.
6AG5
4TH PIX I-F AMPL.
TO PLATES OF 6J6 MIXER
C18 270
PRIMARY OF T2
R14 1000
C9 1500
B+
R115 10K
BIAS
135V
R116 39
C116 270
PRIMARY OF T103
R118 1000
B+
R120 10K
BIAS
135V
C120 1500
R121 39
C123 270
PRIMARY OF T104
R123 1000
B+
R125 47K
BIAS
135 V
R127 2700
C127 1500
B+
C128 270
L183
99V
R134 5600
127V
C135 1500
PRIMARY OF T105
R133 50
C131 82
B+
C134 270
TO CATHODE OF 6AL5 VIDEO DETECTOR
L185

Courtesy RCA

FIG. 6-7.—Simplified schematic diagram of the video i-f section of RCA model 630TS in which the coupling between stages is simple inductive-capacitive coupling.

veloped across each coil is coupled to the following video amplifiers through these coupling capacitors. In order to understand how these video i-f stages contribute to the final resultant curve, we will study the circuit from the input to the first video i-f amplifier.

The signal output from the 6J6 mixer tube consists of a number of different signals. Besides the difference frequency between the local oscillator and incoming r-f signal, there is also their sum, the individual r-f and oscillator signals themselves, and numerous harmonics of all these signals. The primary of *T2* is made selective to the difference frequency between the incoming r-f and oscillator signal. The response of this circuit is single peaked, but broad enough to pass the complete band of the produced i-f signal. However, due to the single-peaked effect, not all signals are equally amplified. The primary of *T2* is loaded down by the 1000-ohm plate load resistor *R*14 and 10,000-ohm grid leak resistor *R*115. This coil circuit, and some of the others, can be represented by the simple network of Fig. 6-8. Capacitor *C* represents the over-all capacitance in the circuit (including the total shunting capacitances, the distributed capacitance of the coil, and the effect of the coupling capacitors) and it is essentially in resonance with the coil. The other components are the same as in Fig. 6-7. The loading down of this resonant circuit by the resistors reduces the *Q* and makes the response less sharp than it would be without this loading effect.

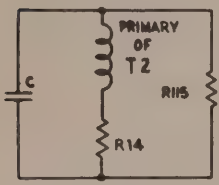
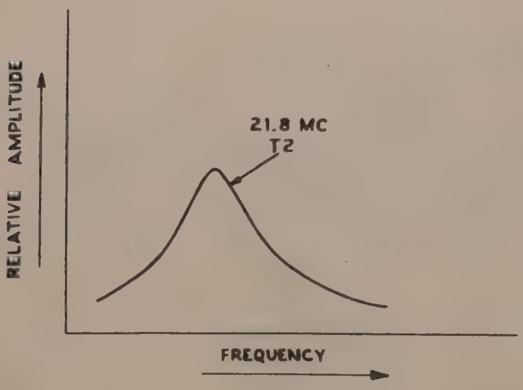


Fig. 6-8.—In this simple network representing the primary of *T2*, *C* is the over-all capacitance in the circuit; *R*115 is the grid-leak resistor and together with *R*14 the load resistor, reduces the *Q* and makes the response less sharp than it otherwise would be.

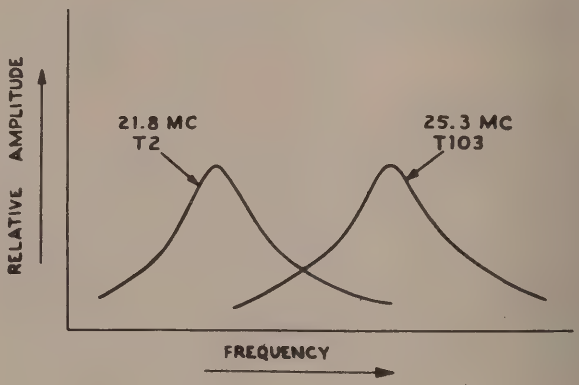


Courtesy RCA

Fig. 6-9.—Response curve for the first video i-f stage which is peaked at a frequency of 21.8 mc.

The response curve for this first video i-f stage is shown in Fig. 6-9. It is peaked at a frequency of 21.8 mc.

In this receiver, the oscillator is so designed that it will track above the incoming r-f signal; hence a difference frequency of 25.75 mc results for the video i-f carrier and 21.25 mc for the sound i-f carrier. Since the mixing process produces an i-f video carrier of 25.75 mc, how can the 21.8-mc resonant circuit be selective enough to pass this frequency to the video amplifiers? Although the primary of *T2* is peaked to 21.8 mc, its response curve is broad enough to pass signals up to 25.25 mc, and somewhat beyond this frequency, to the following stages for amplification. This is a very important fact about stagger-tuned circuits and must be understood in order to comprehend fully how the shaping of the final over-all response curve is possible.



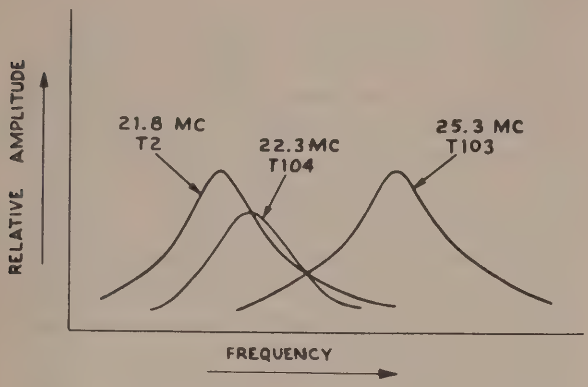
Courtesy RCA

Fig. 6-10.—The response curve of *T*103 is shown to be essentially the same as that of *T2*. *T*103 is also loaded down by plate and grid-leak resistors, but is tuned to a peak resonant frequency of 25.3 mc.

The signal input to the first video 6AG5 i-f amplifier (also called a pix i-f amplifier) contains all the frequencies necessary for proper reception. However, these frequencies are not of equal amplitude due to the response characteristic of the primary of *T2*. This first 6AG5 tube does not change the relative amplitude of the signal input.

The primary of *T*103 is loaded down by plate and grid-leak resistors of the same value as before, and so the shape of its response curve is essentially the same as that of Fig. 6-9. However, this coil is tuned to a peak resonant frequency of 25.3 mc. This response curve and that of *T2* are drawn in Fig. 6-10. The amplitude of this curve is essentially determined by the gain of the second video i-f amplifier. The output of the second video i-f amplifier is impressed across the next tuned circuit.

The primary of *T*104 is tuned to a frequency of 22.3 mc and it, also, has a single-peaked response curve. The shape of this response curve differs from the previous curves because the circuit is still further loaded down. Even though the value of the plate-load resistance has not changed, the 4700-ohm grid-leak resistor *R*125 is much smaller than the 10,000-ohm resistor used in the previous instances. Since the grid-leak resistor is effectively in *shunt* with the tunable coils, then the greater the resistance, the higher the effective coil *Q*, and the lower the resistance, the lower the coil *Q*. This means that the 4700-ohm grid-leak resistor *R*125 effectively lowers the *Q* of the circuit, reducing the gain of this i-f stage but at the same time making this stage much broader than the others. The amount of reduction in gain and increase in bandwidth is readily noticeable when the response of this network is compared to that of the previous two. All three curves are illustrated in Fig. 6-11, where it is seen that the 22.3-mc peak response curve is reduced in amplitude but broader than the other two. The third 6AG5 video amplifier contributes to the gain of this stage. The output of this amplifier is effectively impressed across the *L*183 network as seen in Fig. 6-7.

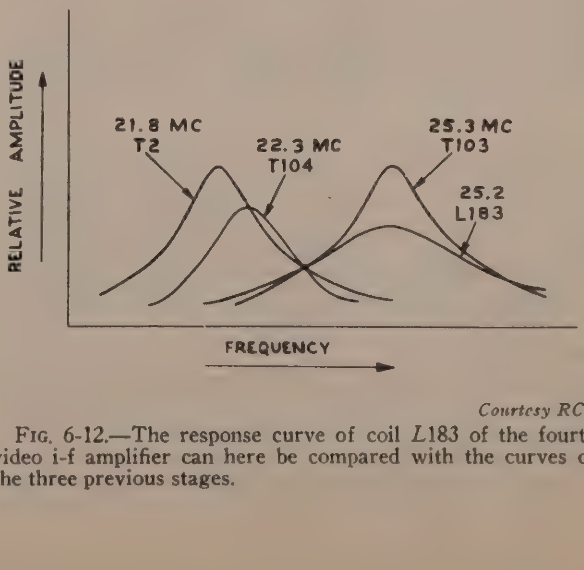


Courtesy RCA

FIG. 6-11.—The primary of *T*104 has the same single-peaked response of *T*2 and *T*103; however, the shape of the curve differs from the others because it is loaded down still further. Its much smaller grid-leak resistor *R*125 is only 4700 ohms compared with the 10,000-ohm resistors used with the other primaries.

The resonant circuits that follow the third video i-f amplifier are arranged somewhat differently from the previous ones. Coil *L*183 is placed in the grid circuit of the fourth 6AG5 video i-f amplifier and thus no grid-leak resistor is used in this circuit. This coil provides the d-c return path for the grid in this fourth video i-f amplifier. The reduction in coil *Q* of this circuit is primarily brought about by the 2700-ohm plate load resistor *R*127 of the third video i-f amplifier. The plate load resistor is effectively in parallel with *L*183 through the coupling capacitor *C*128. However, due to the absence of a grid-leak resistor and to the value of *R*127, the *Q* of the coil is reduced to a greater degree than in the other circuits. This is a necessity in order to obtain the correct over-all video i-f response curve. The resonant frequency of coil *L*183 is 25.2 mc, only 0.1 mc lower than that for the primary of *T*103. The fourth video i-f amplifier contributes to the gain of the response curve for coil *L*183. However the gain cannot be too high or too low, but should be of the proper amount to insure the correct over-all video i-f response. The plate and screen voltages on this fourth video i-f tube are so adjusted that the tube contributes the necessary amount of gain for the proper over-all i-f response. For comparison purposes, the response of the *L*183 circuit is illustrated in Fig. 6-12 together with the response curves of the three previous stages.

The output from the last or fourth video i-f amplifier is impressed across *L*185, the last tunable video i-f stage. Coil *L*185 is not actually in the plate circuit of the fourth video amplifier as it would seem in Fig. 6-7 but instead is in the cathode circuit of the video detector which follows. The resonant frequency of this last stage is 23.4 mc, and it is loaded down by two components, the 5600-ohm plate load resistor *R*134 and the following detector circuit. Both components are essentially in parallel with coil *L*185; the detector circuit represents a very low resistive load to this coil, so that in conjunction with *R*134, it appreciably reduces the effective *Q* of the coil. This results in quite a broad-band single-peaked response curve. The peak of the curve is not very well defined as in

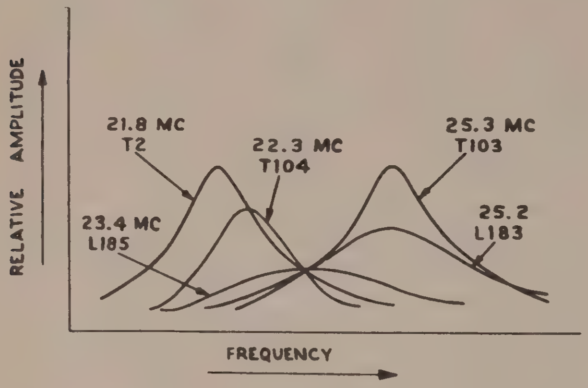


Courtesy RCA

FIG. 6-12.—The response curve of coil *L*183 of the fourth video i-f amplifier can here be compared with the curves of the three previous stages.

the other curves, and the decrease in coil Q also reduces the gain of the *L*185 stage quite a lot. However, the reduction in gain and the broad response obtained is so designed that this last video i-f response curve contributes the final touch to the required shaping of the over-all video i-f response. The response curve of the 23.4-mc circuit is illustrated in Fig. 6-13 in conjunction with the other four response curves. Note how broad this latter curve is compared to the others.

All the five video i-f response curves illustrated in Fig. 6-13 contribute their necessary share of bandwidth and gain so that the over-all video i-f response curve that results from the addition of the curves approaches the ideal curve of Fig. 6-1. The over-all curve for the entire stagger-tuned system is illustrated in Fig. 6-14, this curve is fairly flat and covers the necessary bandwidth. The 25.75-mc mark indicates the position of the video i-f carrier. The broadness of the response of the last video i-f stage is the most helpful in making the over-all curve fairly flat. Without this last video i-f response, the center of the over-all response curve would have quite a dip in its center. The sharpness of the left-hand side of the curve in Fig. 6-14 is due to the traps employed in the video i-f stages.



Courtesy RCA

FIG. 6-13.—The response curve for *L*185, the last tunable i-f stage, is shown in conjunction with the foregoing four response curves. Note the broadness of this last curve compared with the others.

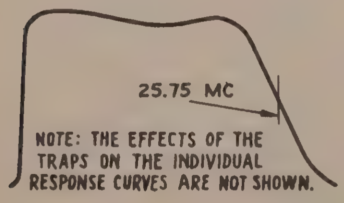


FIG. 6-14.—The over-all curve for the entire stagger-tuned system is shown to be fairly flat and capable of covering the necessary bandwidth. The 25.75 mark indicates the position of the video i-f carrier.

It is interesting to note that in the first three video i-f stages the coils are in the B-plus lead of the plate circuit of the mixer tube, first video i-f amplifier, and second video amplifier respectively, and thus d.c. (plate current) flows through each of these coils. The last two video i-f coils (*L*183 and *L*185) are not in any B-plus circuit and thus d.c. does not flow through these coils.

The Need for Sound and Video I-F Traps

In this chapter we have mentioned the importance of the sound and video i-f traps in shaping the final over-all video i-f response curve. These traps, although helping shape the over-all video i-f response curve, are essentially inserted to prevent interference from undesired sound and video signals. The possibility of interference from different signals will be considered before we analyze the different traps as used in present-day receivers.

If any sound i-f signal passes through the video i-f stages, it will probably be detected and thus cause interference in the reproduced picture. One type of sound interference may be recognized by a series of horizontal bars across the screen of the picture tube.

It may not be readily evident how a sound f-m i-f signal can be detected by the video detector, which is usually a simple diode detector and as such responds only to a-m changes in its input signal. This detection is possible because of the position of the interfering sound f-m signals on the video i-f response curve. These signals lie on the outskirts of the response curve, where the shape of the curve has a definite slope. Due to the shape of the slope, the f-m signal will be changed to one that also varies in amplitude; this amplitude variation is essentially at the rate of the audio signal modulating the f-m wave. Since the new signal is varying in amplitude, it can thus be detected. This process of changing the f-m signal to one that also varies in amplitude is known as slope detection and was discussed in greater detail in chapter 5.

Besides sound i-f interference, there is also the possibility that the video signals adjacent to the one which is desired may be passed and detected, causing a blurred and incoherent pattern on the screen of the picture tube. These signals can easily cause interference, because without the use of special traps it is quite difficult to obtain the desired shape of the over-all video i-f response curve, particularly at the extreme ends of this curve. The shaping of the sloping right-hand side of the curve of Fig. 6-1 with the video i-f carrier frequency approximately in the center of the slope is accomplished chiefly by the video i-f transformers and not by the traps. However, the sharp response on the left-hand side of the curve of Fig. 6-1 is due to the traps employed. If no traps were employed, the shape of the response curve might be broad enough (although not at the same level) to fall outside its desired bandwidth. This means that undesired signals falling inside this oversize response can easily be passed through the video i-f circuits.

Since each television channel is only 6 mc wide, the sound i-f carrier of the next lower channel is only 1.5 mc away from the video i-f carrier to which the set is tuned, and only 0.25 mc away from the limits of the channel. In similar fashion, the adjacent video i-f carrier of the next higher channel is 1.5 mc away from the sound i-f carrier to which the receiver is tuned and 1.25 mc away from the upper end of the channel limit. If any of these adjacent-channel signals pass through the r-f and mixer stages, they will be converted to i-f signals similar to the desired signal. The frequency separation between the carriers will remain the same even after they are converted into i-f signals.

Remember that before frequency conversion the video *r-f carrier* is 4.5 mc lower than the sound *r-f carrier.* After frequency conversion the video *i-f carrier* is 4.5 mc higher than the sound *i-f carrier,* due to the heterodyne oscillator in the receiver tracking above the incoming signal. Thus the video i-f carrier will fall on the high end of the over-all i-f response, and the sound i-f carrier will fall on the low side of the over-all i-f response. In a similar fashion, the adjacent video i-f carrier of the next *higher* channel will appear on the low side of the over-all i-f response, near the position of the sound i-f carrier.

In order to visualize how undesired signals can pass through, let us refer to Fig. 6-15. The solid line curve represents a normal over-all video i-f response curve after proper alignment of the video i-f stages. The shape of this curve may differ somewhat for various receivers due to the type of coupling between stages and the number of different traps employed. For the sake of simplicity, it is shown as having an ideal flat top. Points *A* through *D* represent different frequencies. Point *C* represents the video i-f carrier to which the receiver is tuned. Point *B* represents the accompanying sound i-f carrier to which the receiver is also tuned. If the video i-f carrier, point *C,* is at 25.5 mc, the sound i-f carrier point *B,* being 4.5 mc away, is at a frequency of 21 mc. Point *A* represents the adjacent video i-f carrier and point *D* the adjacent sound i-f carrier. The frequency separation between these i-f carriers is indicated on the drawing. The long solid vertical lines indicate the 6-mc limits of channel bandwidth. The frequency indications of the adjacent sound, point *D,* and video, point *A,* i-f carriers are calculated on the basis of the frequencies assumed for points *B* and *C.*

If traps were not employed, the response curve of the over-all video i-f network of Fig. 6-15 would not take on the shape indicated by the solid line part of the curve but would be different at the ends, probably taking on a shape at these extreme points as indicated by the dashed lines. Upon a quick inspection of the curve, with regard to the shaping due to dashed lines, it is readily seen that points *A, B,* and *D* all fall within the dashed portion of this curve. All the signals represented by these three points are undesired, but they would be detected because all fall within the region of response curve. If the amplitude of the dashed lines could be considerably reduced, little or no interference would result. This reduction is accomplished by the use of special traps which absorb or attenuate these undesired signals. The shape of the response curve with traps employed at frequencies represented by points *A, B,* and *D* will be such that, instead of following the dashed line, it will take on the shape of the solid line as indicated in Fig. 6-15. The amplitude of the solid line curve at points *A, B,* and *C* is seen to be negligible, compared with the over-all height of the curve, and the signals at these points will therefore not be able to pass through the video i-f networks.

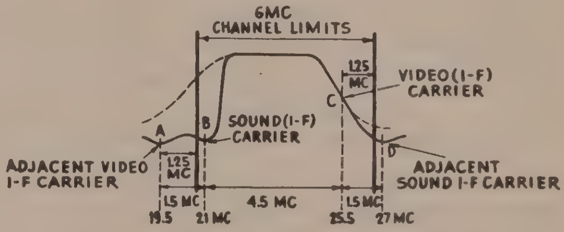


FIG. 6-15.—A normal over-all video i-f response curve with an ideal flat top is shown by the solid-line part of this curve. If traps were not used, the curve would probably end as indicated by the dashed lines, allowing undesired signals at A, B, and D to pass. Reduced in amplitude through the use of traps, their interference becomes negligible.

Practical I-F Traps

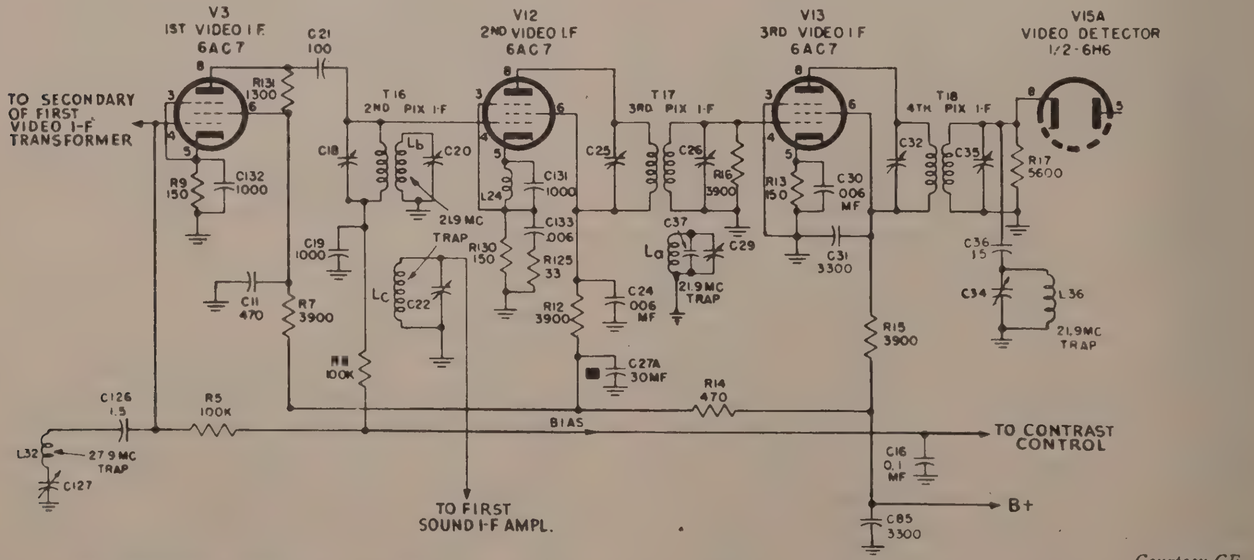
A trap is usually a resonant circuit sharply tuned at the frequency to be suppressed. The resonant trap effectively either absorbs energy or causes a loss in gain at its resonant frequency. Either a series or a parallel resonant circuit can be used as a trap. In the following analysis of different video and sound i-f traps, it should be remembered that a series resonant circuit offers a minimum impedance (which is purely resistive) at resonance and a parallel resonant circuit offers a maximum impedance (also purely resistive) at resonance.

Not all the television receivers employ the same number of i-f traps. The number of traps employed (for different frequencies) is determined by how wide the desired video i-f pass band must be. In those receivers that use a small picture tube, where loss of high video frequencies is not too critical, the pass band is often small with consequent reduction in bandwidth at the high video frequency end. A narrower bandwidth may make unnecessary the use of a trap for the adjacent video i-f carrier or even, perhaps, for the sound i-f carrier of the channel in question. In most cases, however, two or three traps are employed and in special cases four traps may be employed. The *number of traps* as referred to in the discussion so far means those traps that have different resonant frequencies and not to the actual numerical amount of traps used. Some receivers may employ four individual traps but two may be resonant to the adjacent sound i-f and the other two to the accompanying sound i-f. Two traps of the same frequency are used to give better insurance of rejection at the frequencies to which they are tuned.

In the GE model 802 five individual traps are employed, four are tuned to the accompanying sound i-f signal and one to the adjacent sound i-f signal. A schematic of the video i-f section of this receiver, including all these traps, is shown in Fig. 6-16. Two of the traps are series resonant circuits and the other three are parallel tuned circuits. The video i-f carrier of this set is equal to 26.4 mc, and the accompanying sound i-f carrier, being 4.5 mc less, is equal to 21.9 mc. Thus the four traps to suppress the accompanying sound i-f are tuned to 21.9 mc. Since the adjacent sound i-f carrier is 6 mc higher (the distance of a channel width) than the accompanying sound i-f carrier, or 1.5 mc away from the video i-f carrier, it is at a frequency of 27.9 mc. Thus the single adjacent sound i-f trap is tuned to 27.9 mc.

This adjacent sound i-f trap appears in the grid circuit of the first video i-f amplifier (tube V3) and is a series tuned circuit. It consists of capacitor *C*126 in series with coil *L*32 and variable capacitor *C*127 which helps tune the trap to the 27.9-mc resonant frequency. This trap, as all traps, is sharply tuned; that is, it is highly selective to its resonant frequency. A response curve of a trap circuit will show a very sharp peak with steep sides. This is necessary to insure that only the resonant frequency is suppressed and not other frequencies that might fall within its response curve if the curve were broad. For such highly selective circuits, high *Q* coils are needed. This adjacent sound trap offers a low impedance (due to series resonance) to the 27.9-mc frequency and, since it is effectively wired across the first video i-f amplifier, it will offer a ready path to ground for 27.9-mc signals and they will thus not be amplified by tube *V*3.

The other series resonant circuit, which is tuned to 21.9 mc, the accompanying sound i.f., is situated in the cathode circuit of the video detector *V*15*A*. This trap consists of capacitor *C*36 in series with the parallel combination of *L*36 and variable capacitor *C*34. This circuit, although perhaps this is not readily evident, is essentially in series resonance at 21.9 mc. The com-


V3
1ST VIDEO I F.
6AC7
C21
100
V12
2ND VIDEO I.F
6AC7
V13
3RD VIDEO I F
6AC7
V15A
VIDEO DETECTOR
1/2-6H6
TO SECONDARY OF FIRST VIDEO I-F TRANSFORMER
R131
1300
T16
2ND PIX I-F
T17
3RD PIX I-F
T18
4TH PIX I-F
C18
Lb
C20
C25
C26
R16
3900
C32
C35
R17
5600
R9
150
C132
1000
L24
C131
1000
R13
150
C30
.006
MF
21.9 MC
TRAP
C133
.006
C37
La
C29
C31
3300
C36
15
C19
1000
R130
150
R125
33
21.9 MC
TRAP
C34
L36
21.9MC
TRAP
Lc
C22
C24
.006
MF
C11
470
R7
3900
R12
3900
R15
3900
100K
C27A
30MF
R14
470
C126
1.5
R5
100K
BIAS
TO CONTRAST CONTROL
L32
27.9 MC
TRAP
C127
C16
0.1
MF
B+
TO FIRST SOUND I-F AMPL.
C85
3300

Courtesy GE

Fig. 6-16.—Schematic diagram of the video i-f section of the GE model 802 showing the five individual traps employed in this receiver.

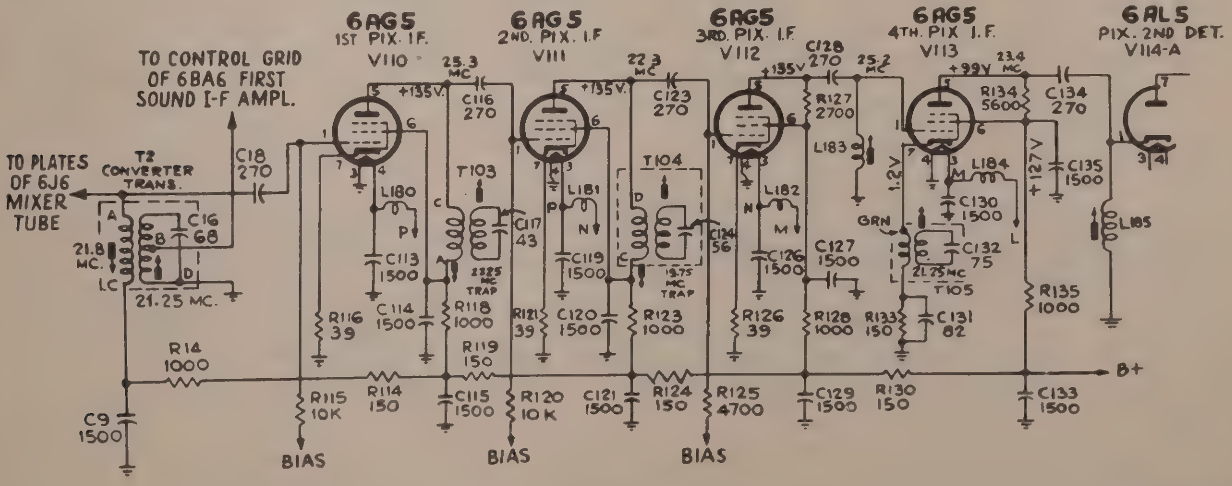
bination of the two capacitors $C36$ and $C34$ in conjunction with $L36$ makes the circuit somewhat sharper in its tuning. Its impedance is a minimum and offers a ready path to ground to the 21.9-mc accompanying sound i-f signal that may enter the detector stage. Thus we see that this trap, as well as the one discussed previously, reduces the gain of the over-all video i-f response at the frequencies to which the traps are tuned.

The 21.9-mc trap used in conjunction with the third video i-f transformer $T17$ works on the principle of absorption. This trap is a parallel resonant circuit which consists of coil L_a in parallel with the fixed capacitor $C37$ and variable capacitor $C29$. The latter capacitor is used to tune the trap to the proper resonant frequency of 21.9 mc. Coil L_a is a tertiary winding on transformer $T17$ and together with its parallel capacitors presents a highly selective tuned circuit at its tuned 21.9 mc. Due to its high selectivity and coupling to $T17$ it absorbs energy at 21.9 mc from the video i-f signal coupled through $T17$. The gain of the video i-f response curve is thus effectively decreased at the tuned frequency of this trap, that of the accompanying sound i.f. The absorbed energy is dissipated in the inherent resistance of the coil.

Another such absorption trap for 21.9 mc is used in conjunction with the video i-f transformer $T16$. This trap consists of coil L_b in parallel with trimmer capacitor $C20$. It is also tuned to the accompanying sound i.f., namely 21.9 mc, and functions in the same manner as the other absorption trap. Below this 21.9-mc trap is another 21.9-mc trap consisting of coil L_c in parallel with trimmer capacitor $C22$. The high side of the coil is connected to the control grid of the first sound i-f amplifier and the low side is grounded. At 21.9 mc this trap offers a maximum impedance and thus, besides causing a loss in the video i-f response at this frequency, a maximum voltage at a frequency of 21.9 mc appears across this circuit and is applied to the first sound i-f amplifier. The selectivity of this latter 21.9-mc trap circuit is such that it will pass the sound i-f signal with its complete bandwidth so that proper reproduction of the audio modulating frequencies will be obtained.

Four traps are employed in the RCA model 630TS. Two traps are employed for the accompanying sound i-f signal, one trap for the adjacent sound i-f signal, and the fourth trap for the adjacent video i-f signal. Since the desired video i-f carrier is 25.75 mc, the accompanying sound i-f carrier, being 4.5 mc less, is equal to 21.25 mc. The adjacent sound i-f carrier is therefore 27.25 mc, and the adjacent video i-f carrier is 19.75 mc. The video i-f circuit for the RCA 630TS receiver is illustrated in Fig. 6-17. The secondary circuit of the converter transformer $T2$ is a parallel resonant circuit tuned to 21.25 mc, the accompanying sound i-f signal. This resonant circuit acts as an absorption trap to the accompanying sound i-f signal, similar to the L_cC22 trap circuit of Fig. 6-16. This trap coil is tapped at point B, and some of the sound i-f signal is taken from the circuit and fed to the control grid of the 6BA6 first sound i-f amplifier.

The parallel tuned secondary circuits of transformers $T103$ and $T104$ both act as sharply tuned absorption traps to the signal to which they are tuned. The secondary of transformer $T103$ is tuned to 27.25 mc and acts as a trap to the adjacent sound i-f signal.


6AG5
1ST PIX. IF.
V110
6AG5
2ND. PIX. I.F
V111
6AG5
3RD. PIX. I.F
V112
6AG5
4TH. PIX I.F.
V113
6AL5
PIX. 2ND DET.
V114-A
TO CONTROL GRID OF 6BA6 FIRST SOUND I-F AMPL.
TO PLATES OF 6J6 MIXER TUBE
T2 CONVERTER TRANS.
A
B
D
LC
21.8 MC.
21.25 MC.
C16 68
C18 270
C9 1500
R14 1000
25.3 MC
+135V.
C116 270
L180
P
C113 1500
T103
C
C117 43
27.25 MC TRAP
R116 39
C114 1500
R118 1000
R119 150
R115 10K
R114 150
C115 1500
BIAS
22.3 MC
C123 270
L181
N
C119 1500
T104
D
E
C124 56
19.75 MC TRAP
R121 39
C120 1500
R123 1000
R120 10K
C121 1500
R124 150
R125 4700
BIAS
C128 270
25.2 MC
R127 2700
L183
L182
M
C126 1500
C127 1500
R126 39
R128 1000
C129 1500
R130 150
1.2V
GRN
+99V
23.4 MC
R134 5600
C134 270
L184
C130 1500
C132 75
21.75 MC
T105
L
R133 150
C131 82
+127V
C135 1500
R135 1000
C133 1500
L185
B+
BIAS

Courtesy RCA

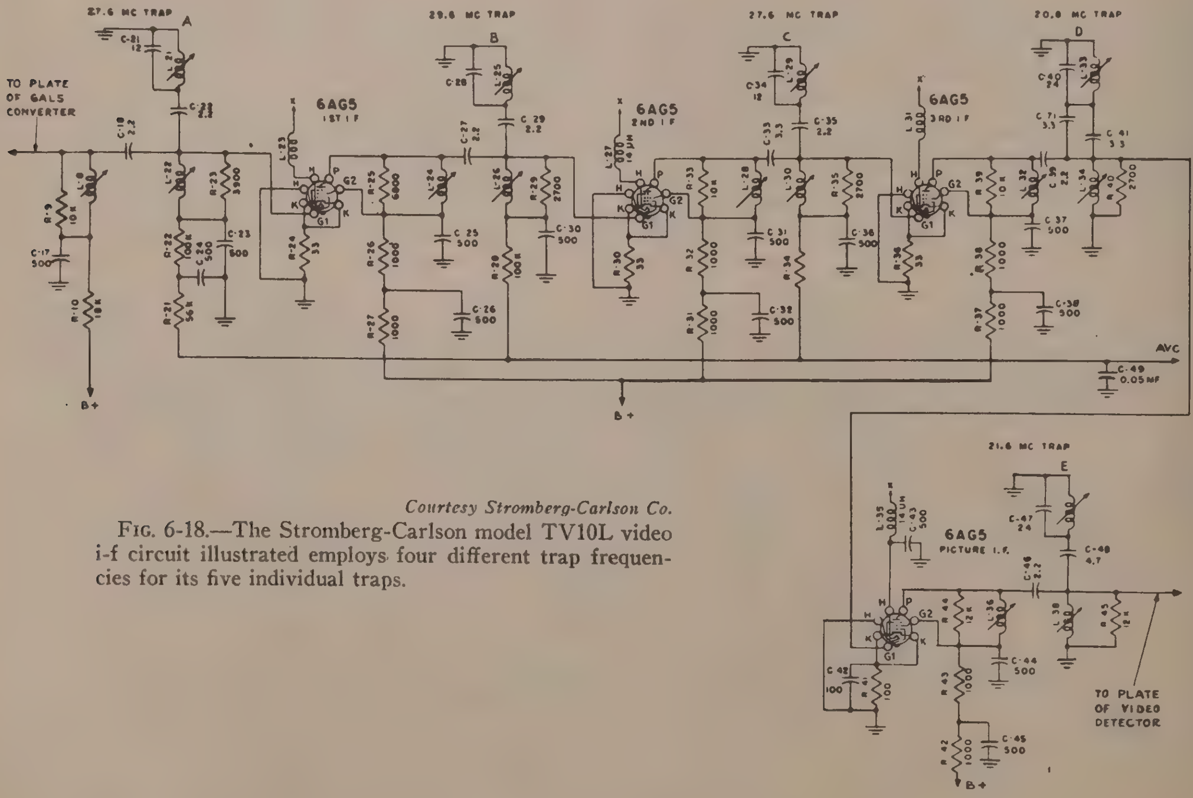
Fig. 6-17.—The video i-f circuit for the RCA 630TS receiver. Note that the secondary circuit of the converter transformer $T2$ is a parallel resonant circuit acting as a sound absorption trap similar to the L_cC22 trap circuit of Fig. 6-16.

The secondary of transformer *T*104 is tuned to 19.75 mc and acts as a trap to the adjacent video i-f signal. The fourth trap in this receiver is located in the cathode circuit of the fourth video i-f amplifier. This trap is the secondary circuit of transformer *T*105 and is parallel resonant to 21.25 mc, the accompanying sound carrier i-f signal.

The primary of *T*105 is in series with *C*131 and forms a series resonant circuit at 23.4 mc, the frequency to which coil *L*185 is tuned. Since a series resonant circuit offers a minimum amount of impedance, the cathode of tube *V*113 is effectively grounded at the frequency to which the primary circuit of *T*105 is resonant. This permits the tube to operate at a gain at this frequency (23.4 mc). The parallel resonant frequency of the 21.9-mc trap has a high impedance and thus reflects a high impedance into the primary of *T*105 at this frequency. This means that the cathode circuit has appreciable impedance, and an r-f voltage drop appears across this circuit at the frequency of 21.9 mc. This r-f voltage drop causes degeneration in the tube to the extent where the tube does not contribute any appreciable gain at this 21.9-mc frequency.

In the Stromberg-Carlson Model TV-10L (series 10) five individual traps are employed, with two tuned to the same frequency, and the other three to different frequencies. This means that we have to deal with four different trap frequencies. In the previous circuits, the maximum number of trap frequencies employed was three, that of the accompanying sound i-f carrier, the adjacent sound i-f carrier, and the adjacent video i-f carrier. In this receiver all these three types are employed, plus an additional one.

Let us examine the video i-f circuit of this receiver which is illustrated in Fig. 6-18. All the video i-f coupling networks essentially consist of a tuned circuit in the plate of one tube and another tuned circuit in the control grid of the next. These networks are coupled together by small value capacitors. Across the grid of each i-f amplifier and the plate of the video detector appears a series resonant trap. Each trap essentially consists of a capacitor in series with a parallel combination of a coil and capacitor. Although not readily evident from the circuit, the components of these traps are so chosen that they are in series resonance at the frequency indicated next to each trap in Fig. 6-18. (These traps are similar to the series resonant trap consisting of C36-L36-C34 in the cathode circuit of the video detector of the GE


27.6 MC TRAP
A
C-21 12
L-21
TO PLATE OF 6AL5 CONVERTER
C-22 2.2
C-18 2.2
L-18
R-9 10K
C-17 500
R-10 18K
B+
L-22
R-23 3900
R-22 100K
C-24 500
C-23 500
R-21 56K
6AG5 1ST I F
L-23
H
P
G2
K
G1
R-24 33
29.8 MC TRAP
B
C-28
L-25
C 29 2.2
C-27 2.2
R-25 6800
L-24
C 25 500
R-26 1000
R-27 1000
C-26 500
L-26
R-29 2700
C-30 500
R-28 100K
B+
6AG5 2ND I F
L-27 14 μH
R-30 33
R-33 10K
L-28
C 31 500
R 32 1000
R-31 1000
C-32 500
27.6 MC TRAP
C
C34 12
L-29
C-35 2.2
C-33 3.3
L-30
R-35 2700
C-36 500
R-34
6AG5 3RD I F
L 31
R-36 33
R 39 10K
L 32
C 37 500
R 38 1000
R-37 1000
C-38 500
20.8 MC TRAP
D
C-40 24
L-33
C 71 3.3
C 41 3.3
C 39 2.2
L-34
R 40 2700
AVC
C-49 0.05 MF
21.6 MC TRAP
E
C-47 24
C-48 4.7
6AG5 PICTURE I.F.
L-35 14 UH
C-43 500
C-46 2.2
R 44 12K
L-36
C-44 500
L-38
R 45 12K
C 42 100
R 41 100
R 43 1000
R 42 1000
C-45 500
B+
TO PLATE OF VIDEO DETECTOR

Courtesy Stromberg-Carlson Co.

FIG. 6-18.—The Stromberg-Carlson model TV10L video i-f circuit illustrated employs four different trap frequencies for its five individual traps.

model 802 in Fig. 6-16.) Since they are in series resonance, they offer a minimum impedance to ground to the frequency for which each one is tuned. Traps *A* and *C* in Fig. 6-18 are resonant to 27.6 mc, which is the adjacent sound i-f carrier. Even though the adjacent video i-f signal has a carrier frequency of 20.1 mc, the 20.8-mc trap (*D*) probably has a response broad enough to attenuate the adjacent video i-f carrier also. Trap *E* is resonant to the accompanying sound i-f carrier of 21.6 mc. Trap *B* is tuned to 29.6 mc, the 20.8-mc trap (*D*) probably has a response in the channel of the adjacent sound i-f signal. This trap will attenuate possible interference from this video i-f signal.

In the Consolidated Television model 2315, all the traps employed are situated in the cathode circuits of the video i-f amplifiers. They function differently from any of the traps discussed so far. A circuit diagram of the video i-f section of this receiver appears in Fig. 6-19. The traps consist of a coil and capacitor in parallel and the frequency to which it is tuned is indicated next to each trap. Although no trap is tuned exactly to the adjacent or accompanying sound i-f carrier or to the adjacent video i-f carrier, they are tuned to frequencies that will help produce the necessary over-all video i-f response. They also provide for the attenuation of signals that might cause interference. They all work on the following principle.

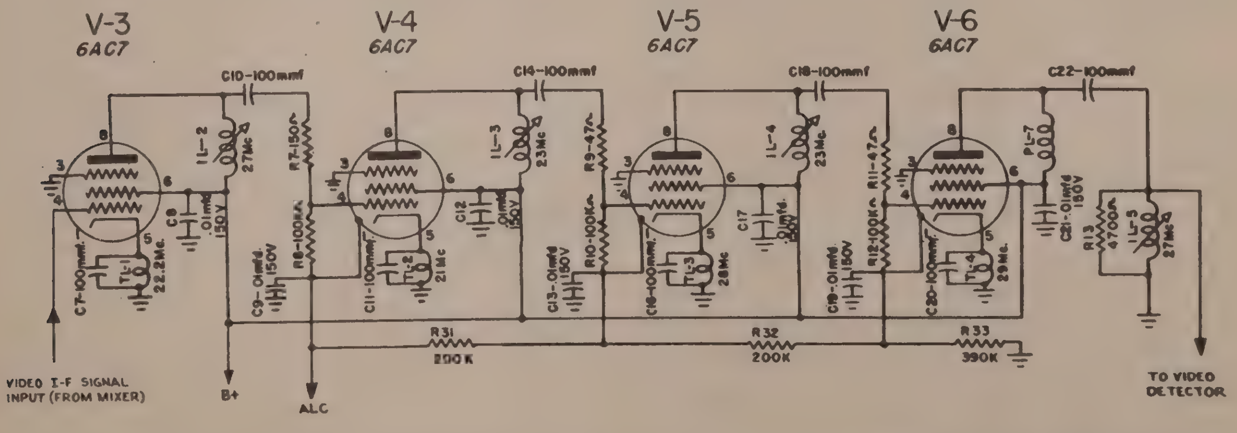
The traps, since they are parallel tuned, offer a maximum impedance to those frequencies for which they are resonant. This means that when the plate r-f current flows through the cathode circuit, a maximum r-f voltage drop will exist across these traps at their resonant frequency. This causes degeneration in each tube at the resonant frequency of the trap connected to it. This means that the tube offers negligible gain at the frequency to which the trap is tuned. The necessary rejection of interference frequencies is thus produced.

Contrast Control

The *contrast control,* sometimes called the picture control, of a television receiver determines the strength of the video signal input to the picture tube. The magnitude of the video signal in conjunction with the picture tube and potentials thereon controls the contrast between the light and dark areas in the reproduced picture. The contrast control, usually a potentiometer, is manually operated. This control is usually inserted either in the video amplifier section preceding the video detector or in the video i-f section of the receiver. Its use in the video amplifier section is discussed in the following chapter. In this section we will study some of the applications of the contrast control in the video i-f section.

The contrast control when used in the video i-f section (as well as the video amplifier section) controls the amount of bias on most of the video i-f amplifiers. By controlling the bias, the amount of i-f gain can be varied to whatever degree is desired, likewise varying the magnitude of the video signal input to the grid of the picture tube. There are many contrast or gain control arrangements used in television receivers. To discuss all the different methods of controlling the gain of the video signal is beyond the scope of this book; however, we will analyze a few of the types used in the video i-f section of present-day television receivers.

In many television receivers, a d-c potential exists across the contrast control, with the positive side of the control grounded. The simple circuit of Fig. 6-20


V-3
6AC7
V-4
6AC7
V-5
6AC7
V-6
6AC7
C10-100mmf
C14-100mmf
C18-100mmf
C22-100mmf
1L-2
27Mc
R7-150Ω
C8
.01mfd
150V
C7-100mmf
T1-1
22.2Mc
C9-.01mfd
150V
R8-100KΩ
1L-3
23Mc
R9-47Ω
C12
.01mfd
150V
C11-100mmf
T1-2
21Mc
C13-.01mfd
150V
R10-100KΩ
1L-4
23Mc
C17
.01mfd
150V
C16-100mmf
T1-3
28Mc
C19-.01mfd
150V
R11-47Ω
R12-100KΩ
PL-7
C21-.01mfd
150V
C20-100mmf
T1-4
29Mc
R13
4700Ω
1L-5
27Mc
R31
200K
R32
200K
R33
390K
VIDEO I-F SIGNAL INPUT (FROM MIXER)
B+
ALC
TO VIDEO DETECTOR

Courtesy Consolidated Television Corp.

FIG. 6-19.—In this Consolidated Television model 2315, all the traps employed are situated in the cathode circuits of the video i-f amplifiers.

illustrates this type of connection. The number of video i-f tubes to which the negative potential is applied is determined by the design of the individual receiver. If the contrast control is advanced to the point where too much bias is applied to the video i-f tubes, their gain will decrease and the video signal input to the picture tube will decrease. This will cause the contrast between the light and dark areas to be less distinguishable by effectively decreasing the intensity of the lighter areas on the screen of the picture. If the contrast control decreases the bias on the video i-f tubes too much, their gain will increase, thereby causing the video signal input to the picture tube to increase. Too great an increase in the magnitude of the video signal will produce a greater contrast between the light and dark areas. This effectively increases the intensity of the lighter areas on the screen of the picture tube to a point where the intermediate grey tones of the picture are lost. The resulting picture will be glaring and incoherent.

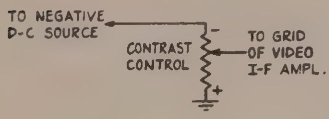

TO NEGATIVE D-C SOURCE
CONTRAST CONTROL
TO GRID OF VIDEO I-F AMPL.
-
+

Fig. 6-20.—This type of contrast control, in which a d-c potential exists across the contrast control and its positive side is grounded, is common in television receivers.

The primary requisite for the operation of the contrast control is a d-c potential across it. This bias is often obtained from a negative d-c source in the power supply. This is done in the Du Mont model RA-102 where the contrast control is part of a voltage dividing network within the 40-volt d-c supply and receives its necessary voltage from this supply. The contrast control is connected to the control grids of the first two video i-f amplifier tubes and varies the gain of these two stages; this in turn varies the video signal input to the picture tube. A schematic diagram of this section of the Du Mont RA-102 receiver is illustrated in Fig. 6-21.

From this diagram we see that the variable arm of the 25,000-ohm contrast control *R*6 is connected to the control grid circuit of the first and second 6AU6 video i-f amplifiers. Variation of this control changes the bias on these tubes and hence their gain. Thus the signal input to the following video i-f amplifiers, video detector, and video amplifiers is effectively controlled by *R*6.

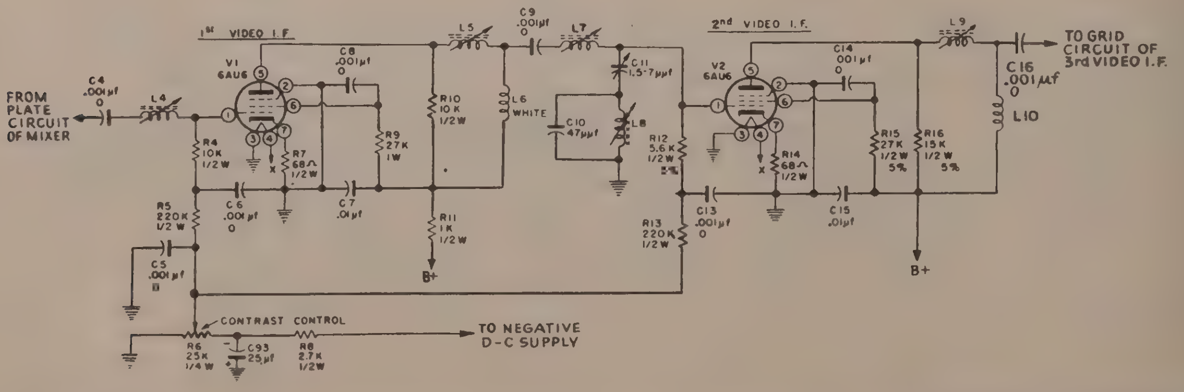
In the GE model 802, the contrast control is in the grid circuit of the first and second video i-f amplifier. This control, unlike the previous one, receives its negative voltage from the grid circuit of another tube in the receiver, the 6AS7G damping tube. The circuit arrangement of the contrast control in conjunction with the damping tube is shown in Fig. 6-22. According to the manufacturer's specifications the voltage measured between grid to ground of the 6AS7 tube is —6.5 volts.

Since the values of *R*120 and *R*121 are very small compared to the addition of resistance *R*90, *R*115, and *R*122, they cause only a negligible drop in the —6.5 voltage. Therefore we can consider the voltage from point *A*, the junction of *R*120 and *R*90 to ground, as being approximately —6.5 volts.

In parallel with this circuit from point *A* to ground is a resistance network consisting of the 10-megohm resistor *R*134 and the 0.5-megohm contrast control *R*108*A*. (The impedance of this network is too high to affect appreciably the operation of the damping tube). Resistor *R*134 and potentiometer *R*108*A* act as a voltage divider network to the —6.5 volts. The voltage appearing across the 0.5-megohm resistance of *R*108*A* is computed as follows:

$$\frac{.5}{10 + .5} \times 6.5 \text{ volts} = 0.31 \text{ volt approximately}$$

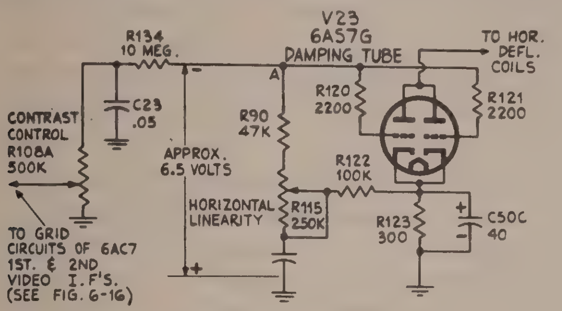
This 0.31 volt is the complete voltage available


1st VIDEO I.F
V1 6AU6
C8 .001µf O
L5
C9 .001µf O
L7
C11 1.5-7µµf
2nd VIDEO I.F.
V2 6AU6
C14 .001µf O
L9
TO GRID CIRCUIT OF 3rd VIDEO I.F.
C16 .001µf O
L10
FROM PLATE CIRCUIT OF MIXER
C4 .001µf O
L4
R4 10K 1/2W
R7 68Ω 1/2W
R9 27K 1W
R10 10K 1/2W
L6 WHITE
C10 47µµf
L8
R12 5.6K 1/2W 5%
R14 68Ω 1/2W
R15 27K 1/2W 5%
R16 15K 1/2W 5%
R5 220K 1/2W
C6 .001µf O
C7 .01µf
R11 1K 1/2W
B+
C13 .001µf O
C15 .01µf
R13 220K 1/2W
B+
C5 .001µf B
CONTRAST CONTROL
TO NEGATIVE D-C SUPPLY
R6 25K 1/4W
C93 25µf
R8 2.7K 1/2W

Courtesy Allen B. Du Mont Labs., Inc.

Fig. 6-21.—The d-c potential across the contrast control is often obtained from a negative d-c source in the power supply as in this Du Mont RA-102 receiver in which the contrast control is within the 40-volt d-c supply.

across the contrast control, and this is the maximum voltage that can be applied to the first and second video i-f amplifier grids. This means that the additional bias applied to these grids will be anywhere from 0 to 0.31 negative volt. This may seem like a very small amount of bias to control the gain of these stages. However if we know the characteristics of the tubes employed as video i-f amplifiers, we will be able to understand how the small contrast voltage can vary their gain. The 6AC7 tubes that are employed as video i-f amplifiers are high transconductance tubes in which the transconductance, or g_m, of the tube varies appreciably with a small change in bias. The fixed bias (cathode bias) on the first 6AC7 video i-f tube is —2.1 volts and on the second video i-f tube it is —2.0 volts. The plate and screen potentials on these 6AC7 tubes are such that at —2 volts the g_m is about 9000 micromhos and at —2.5 volts the g_m is about 7000 micromhos. This range of g_m over the —2 to —2.5 volts small bias change can be considered as being linear. Since the negative voltage across the contrast control is 0.31 volt, the bias on the first i-f tube can be made to vary from —2.1 volts to —2.4 volts and on the second i-f tube from —2.0 volts to —2.31 volts. Thus, even though the —0.31 volt across the contrast control is small, it does cause variations in the bias on the tubes that in turn cause appreciable changes in the g_m of the tubes.



After GE

FIG. 6-22.—In the GE model 802 the contrast control is in the grid circuit of the first and second video i-f amplifier and receives its negative voltage from the grid circuit of the 6AS7G damping tube.

Since the gain of a pentode amplifier is directly proportional to the transconductance of the tube, it is readily seen how this small voltage across the contrast control, *R*108*A* in Fig. 6-22, can vary the gain of the video i-f section and hence the magnitude of the video signal input to the picture tube.

Automatic Gain Control (agc)

If the strength of the signal input to a radio receiver changes, this will produce a change in the output volume of the receiver. To compensate for these changes in output, most radio receivers employ automatic volume control (avc) circuits. In the usual type of avc circuit in radio receivers the received signal is rectified at a certain point in the receiver and applied as a negative d-c signal (avc signal) to the grids of certain tubes, thereby changing their bias and varying their gain. If the input signal increases, the bias on the tube will automatically increase in *proportion* to the increase in input signal, thus keeping the output of the receiver at a constant level. In like manner, if the signal input decreases, the bias will decrease and the gain of the tubes will increase and thus tend to keep the output of the set level.

In television reception the strength of the signal input to the television receiver can likewise change, causing undesired variations in the video signal input to the picture tube. In many television receivers the video i-f signal is rectified and applied as a negative d-c signal, usually to the grid circuits of two or more of the stages preceding the detector. This signal varies the bias and therefore the gain of the stages in much the same way avc does in radio receivers. Although at times this variation is referred to as avc in the video channel of television receivers, it is usually known as *automatic gain control* (agc), or sometimes as automatic level control (alc). There are various different means of obtaining this automatic control voltage, but diode rectification of the video i-f signal is most often used.

A simple arrangement of an agc circuit is shown in Fig. 6-23. The video i-f signal is transformer coupled to the plate of the diode. Since the cathode is grounded, current will flow through the tube only on the positive half cycle of the input signal. The electron current flow will be from cathode to plate, through the plate-load resistor R_2 to ground, then back to the cathode of the tube to complete the d-c path. Due to the direction of this current flow, indicated by the arrow in the drawing, the potential on resistor R_2 will be negative at its plate connection with respect to its ground connection. This negative

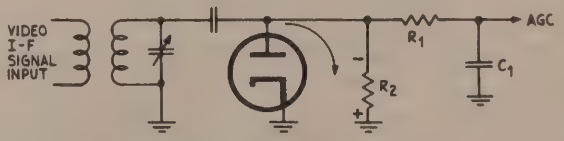
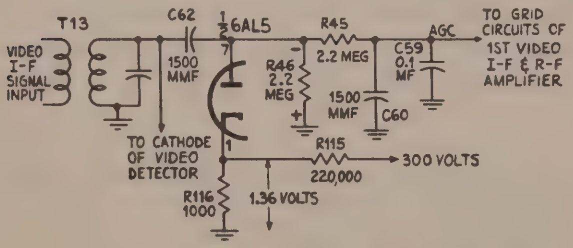


FIG. 6-23.—Due to the direction of the current flow in this simple agc circuit (indicated by arrow), the potential on R_2 will be negative with respect to ground, presenting a ready point for tapping of agc.

part of the load resistor presents a ready point for tapping off agc. However, the current flowing through the tube varies in accordance with the envelope of the modulated input signal; to convert this output signal as closely as possible to a d-c signal and to prevent any i-f voltage from getting to the automatic-gain-controlled tubes, a filter network is employed. This filter is usually a resistance-capacitance network as shown by the R_1C_1 circuit in Fig. 6-23 and tends to level out the varying output signal.

In many cases no agc signal is desired on the grids of the tubes to be controlled until the input signal to the receiver reaches a certain value. This is to make sure that on weak signal inputs there will be no agc signal available and the tubes will thus offer maximum gain at low signal inputs. Such a circuit when used in radio receivers is called *delayed avc*. When used in the video section of television receivers, it is commonly known as *delayed agc* because there is a delay before an agc signal is available.

In the Belmont model 22A21 a 6AL5 duo-diode tube is used, one diode section of which is used for video detection and the other diode for *delayed agc*. The circuit for this agc arrangement is shown in Fig. 6-24. To make sure that the diode will not conduct on weak signal inputs, a positive potential of approximately 1.36 volts is put on its cathode. This potential is obtained from the 300-volt B supply through the voltage divider network consisting of the 220,000-ohm resistor *R*115 and the 1000-ohm resistor *R*116



Courtesy Belmont Radio Corp.

FIG. 6-24.—This arrangement for a delayed agc is used in the Belmont model 22A21. One section of the 6AL5 duo-diode tube is used for video detection and the other for the delayed agc.

Since the complete 300 volts exists across the series combination of *R*115 and *R*116, then that portion of voltage existing across *R*116 alone is as follows:

$$\frac{R116}{R115 + R116} \times 300 \text{ volts}$$

$$\frac{1000}{221{,}000} \times 300 = \frac{300{,}000}{221{,}000} = 1.36 \text{ volts}$$

Thus we see that a positive potential of 1.36 volts exists on the cathode of the tube. A diode functions in such a manner that it conducts only when the potential on its plate is more positive than that on the cathode. Thus for the circuit of Fig. 6-24, the tube will conduct only when the potential on the plate exceeds 1.36 volts. This means that not until the video i-f signal input at the plate of the diode (which is coupled through capacitor *C*62) has a peak signal strength equal to 1.36 volts will the tube conduct. (Remember that the value of 1.36 volts is dependent upon the values of *R*115 and *R*116 being exactly as listed, but since each has a 10-percent tolerance, the potential on the cathode may be somewhat different from that calculated. However, the difference will not be great.)

The tube will conduct only when the positive half cycle of video i-f signal input exceeds a peak value of 1.36 volts. When the tube starts conducting, a d-c voltage drop will exist across the plate-load resistor *R*46 with the plate side of the resistor becoming negative with respect to its grounded side, as shown in Fig. 6-24. The signal across *R*46, after passing through a resistance-capacitance filter network consisting of *R*45, *C*60, and *C*59, is applied as a negative d-c signal to the grid circuits of the first video i-f amplifier and the r-f amplifier of the receiver.

The stronger the video i-f signal input to the diode, the greater will be the current flow and, hence, the greater the voltage drop across *R*46. This means that the bias on the first video i-f tube and the r-f amplifier tube will increase, thus reducing their gain. With a strong signal input, a high (negative) agc signal results, which reduces the gain of the controlled tubes in such a manner that the input video signal to the picture tube is effectively kept at a constant level.

Video I-F Amplifier Tubes

The video i-f amplifier section contributes most of the gain of the video channel of the television receiver. Most of the video i-f amplifier tubes employed have a high transconductance in order to contribute the necessary gain. A large g_m is a fundamental requirement for high gain from pentode amplifiers because the gain is directly proportional to the g_m of the tube. Some of the different tubes employed as i-f amplifiers are the 6AC7, 6AU6, 6AH6, and the 6AG5. The transconductances of these tubes is quite high, their values ranging from about 5000 to 10,000 micromhos.

Besides having high transconductance values, those tubes that are generally used as video i-f amplifiers have sharp cutoff characteristics. The use of sharp cutoff tubes decreases the distortion of the system because their plate current—grid voltage characteristic curves are more nearly linear at the bottom of the curve than those of remote cutoff tubes. If the i-f amplifier is not a sharp cutoff tube, the plate current—grid voltage characteristic curve will have appreciable curvature at its bottom. If the input signal is such that it operates over the lower portion of this curve, distortion will be present in the output of the amplifier. When high transconductance tubes with sharp cutoff characteristics are used, a small change in bias produces a large change in plate current.

In some cases when agc or contrast control voltages are used, it is desired that the tubes have a remote cutoff characteristic. This gives the control voltage a wide latitude of variation in changing the plate current of the tube. Some sharp cutoff tubes exhibit remote cutoff characteristics when a screen-grid resistor is used. By the use of the proper resistor, the d-c potential on the screen grid can be made to vary with the bias on the control grid in such a manner that the tube will exhibit remote cutoff characteristics. Many tube manuals list a typical value of screen resistor to use with sharp cutoff tubes to obtain remote cutoff characteristics.

VIDEO DETECTOR

The tube following the last video i-f amplifier is usually the video detector. The video detector circuits of television receivers are very similar in operation to those employed in a-m radio receivers, because the video i-f signal is an a-m wave and essentially requires the same basic type of detector circuit as any a-m wave. (This detector circuit is often called the second detector when the mixer or converter tube used is called the first detector. In this chapter, however, only the term detector will be used and not second detector.)

In an a-m receiver the detector rectifies the incoming i-f signal, and after passing through special i-f filters, the modulating audio signal corresponding to the amplitude variations of the i-f carrier is recovered. This audio signal is then applied to the necessary audio amplifiers for reproduction in the output of the speaker. In a similar manner, the video detector in a television receiver rectifies the video i-f signal and, after proper filtering, the modulating video signal (including the sync and blanking signals) corresponding to the amplitude variations in the video i-f carrier is recovered. After the proper separation is made, the picture (true video) signals and sync signals are applied to their respective circuits for correct picture reproduction.

The video detector generally employed is of the diode rectifier type. Although the basic operation of this type of circuit is essentially the same as that employed in a-m receivers, the design of the circuit is very different. In the design of video detector circuits such factors as the required polarity of the video signal on the picture tube, the loading on the detector, and the frequencies involved must be taken into account. All these factors will be considered in the following sections.

Phasing of the Picture Signal

It was mentioned that the required polarity of the video signal input to the picture tube is a determining factor in the design of the detector circuit. In other words, the video signal input to the picture tube has to be correctly phased for the tube to reproduce the picture as taken at the studio. If we were to examine the schematics of a number of television receivers, we would note some very interesting details in the video channel following the last video i-f amplifier section. Following this last video i-f amplifier is the usual video diode detector, and following that the video amplifiers. In most cases the video signal is fed into the control grid of the picture tube, although in some instances the signal may be fed to the cathode of the picture tube. In either case the *effective* signal on the grid of the picture tube should be a negative video signal.

In those receivers where the video signal is fed directly into the control grid of the picture tube from the plate circuit of the last video amplifier and where an *odd number* of video amplifiers is employed, the video i-f signal input to the diode detector will be coupled to the *plate* of the diode. When an *even number* of video amplifiers is employed, the video i-f signal input to the detector will be coupled to the *cathode* of the diode. When the video signal is fed to the cathode of the picture tube, the diode detector connections just discussed are reversed when the same number of video amplifiers are used. These different connections are necessary because a negative video signal is essential at the grid of the picture tube to insure proper reproduction of the picture.

In the reproduction of an audio signal, the phase of the signal input to the speaker is not important because faithful reproduction of the audio signal does not depend upon the polarity of the audio signal input to the speaker. This means that the ear is not sensitive to phase changes in the audio signal. However, a different situation exists in television regarding the polarity of the video signal input to the picture tube, because the human eye is more sensitive than the ear. If the polarity is reversed, the reproduced picture will be a *negative* of that at the studio. In other words, the black parts of the picture will be reproduced white and the white parts reproduced black.

In chapter 1 we stated that negative transmission is used for television. This means that in the video a-m signal the positive peaks of the video modulating signal will represent the black portions of the signal and the negative peaks the white parts of the picture. A video modulated signal is shown in Fig. 6-25 together with the modulating signal. For the sake of discussion the video signal is shown as a sine wave. In the video modulating signal, Fig. 6-25 (A), the positive peaks are the black regions of the signal and the negative peaks the white regions. In the modulated signal, Fig. 6-25 (B), the peak represents the black regions and the troughs represent the white region. The sync pulse and blanking signals are also shown superimposed upon the carrier signal. When the modulated signal is applied to the video detector, it will be rectified and the undesired r-f (carrier) will be properly filtered out. The signal that remains after this process is the modulating component, which consists of the video signal and the superimposed sync and blanking signals. This composite signal is then fed through the video amplifier system. After amplification and then d-c reinsertion, the composite video signal is ready to be fed into the picture tube.

When the control grid of the picture tube has its bias *effectively* decreased (that is, made less negative), a stronger beam of electrons will flow in the tube, causing a brighter image upon the screen of the picture tube. When the polarity on the control grid of the picture tube is reversed, the positive peaks of the video signal, which represent the black part of the transmitted picture, would make the grid less negative and the reproduced image would be brighter. That is, the black part of the transmitted picture would effectively be reproduced white. In like manner the negative peaks of the same signal, which represent the white parts of the transmitted signal, would be reproduced black.

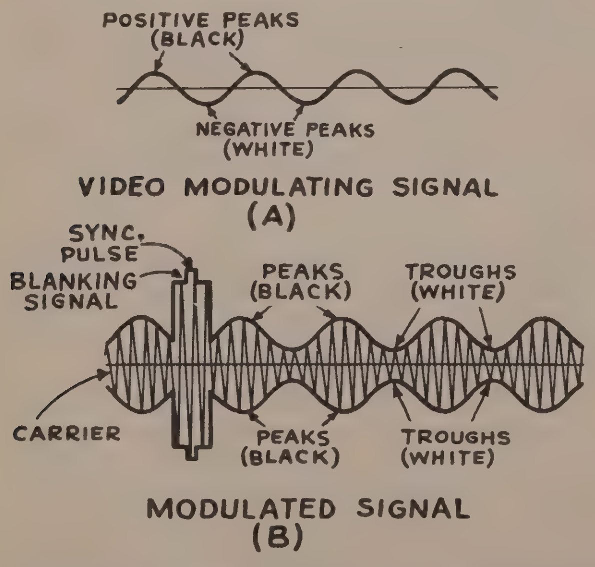


Fig. 6-25.—How negative transmission is used for television is here illustrated by representing the video signal as a sine wave. In (A) the positive peaks are the black regions of the picture and the negative peaks are the white regions. In (B) the peaks represent the black regions and the troughs the white regions.

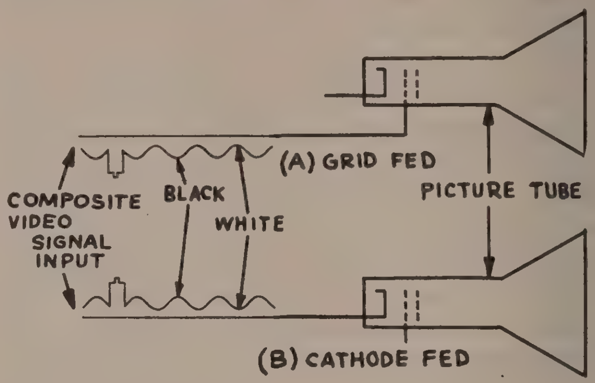


Fig. 6-26.—For correct operation of the picture tube the black parts of the video signal should drive the grid more negative and the white parts, less negative. Depending upon the polarity of the signal, it may be fed to either the grid (A) or the cathode (B) of the tube.

Besides the composite video signal reproducing a negative picture as just described, the sync and blanking pulses of the signal would result in bright retraces across the screen. In some cases, perhaps, the sync and blanking signals might be strong enough to drive the grid positive and possibly cause the cathode of the picture tube to burn out.

For correct operation of the picture tube, the black parts of the video signal should drive the grid more negative and the white parts less negative, as illustrated in Fig. 6-26. In this manner the brighter parts will cause more electrons to flow from the cathode of the picture tube and the blacker parts less electrons. In Fig. 6-26 (A) the polarity of the video signal is such that it lies below the reference line. Thus the black parts of the signal are more negative and will increase the bias on the grid and therefore reproduce black images on the screen. The white parts of the signal are less negative and will reduce the bias on the grid and, therefore, cause more electrons to flow, producing white images on the screen.

If the video signal is fed into the cathode of the picture tube, the phasing of the signal must be opposite to that when fed into the grid. This is illustrated in Fig. 6-26 (B) where the video signal is seen to lie above the reference level. The black parts of the signal make the cathode more positive with respect to the grid, which means that the grid becomes more negative with respect to the cathode, thus decreasing the number of electrons hitting the screen. The white parts of the signal make the cathode less positive with respect to the grid, thus effectively decreasing the bias on the grid and increasing the number of electrons hitting the screen. Consequently, feeding the video signal to the cathode with the proper degree of polarity, as shown in Fig. 6-26 (B), will result in a correct picture on the screen. From this analysis we can readily see the necessity for the correct polarity of the video signal when fed either to the grid or cathode of the picture tube. In parts (A) and (B) of Fig. 6-26, the level of the blanking signal is such that it will drive the picture tube beyond cutoff.

It was stated previously that the video i-f signal input to the detector may be connected to either the plate or cathode of the diode detector, depending upon the number of video amplifiers employed and the type of connection to the picture tube. In order to visualize the various possible types of circuit arrangements in the video section when the video signal output is fed to the picture tube from the plate circuit of the last video amplifier, let us refer to the simplified circuits of Fig. 6-27. If the video i-f signal is fed into the plate of the diode detector, then after rectification and proper filtering, the modulating component that is fed into the following video amplifiers will lie above the zero reference line, or in the so-called positive region, as indicated in Fig. 6-27 (A). With this detector arrangement, an odd number of video amplifiers is needed to produce the correct polarity of signal input to the grid of the picture tube; and an even number of video amplifiers is required when the video signal is fed to the cathode of the picture tube.

On the other hand when the video i-f signal input to the diode detector is to the cathode of the diode, as shown in Fig. 6-27 (B), the polarity of the output signal is reversed from that in Fig. 6-27 (A). The use of even and odd numbers of video amplifiers should be reversed from that of part (A) to give the correct polarity of signal input to the cathode or grid of the picture tube. This is indicated in Fig. 6-27 (B). Usually either one or two video amplifiers are employed.

Detector and Amplifier Action

It may not be readily evident how and why the phasings of the video signal, as indicated in Fig. 6-27 (A) and (B), come about. The phasing of the video signal depends upon the action of the diode rectifier and upon the inherent phase inversion qualities of amplifying vacuum tubes. The video signal input to the diode detector is a modulated wave as shown in Fig. 6-25 (B), assuming a sine wave video modulating signal. This input signal is also indicated in Fig. 6-28 and is fed into the plate of a diode rectifier in (A) and into the cathode in (B). That part of the video modulated signal that lies above the reference line is considered to be in the positive or plus (+)

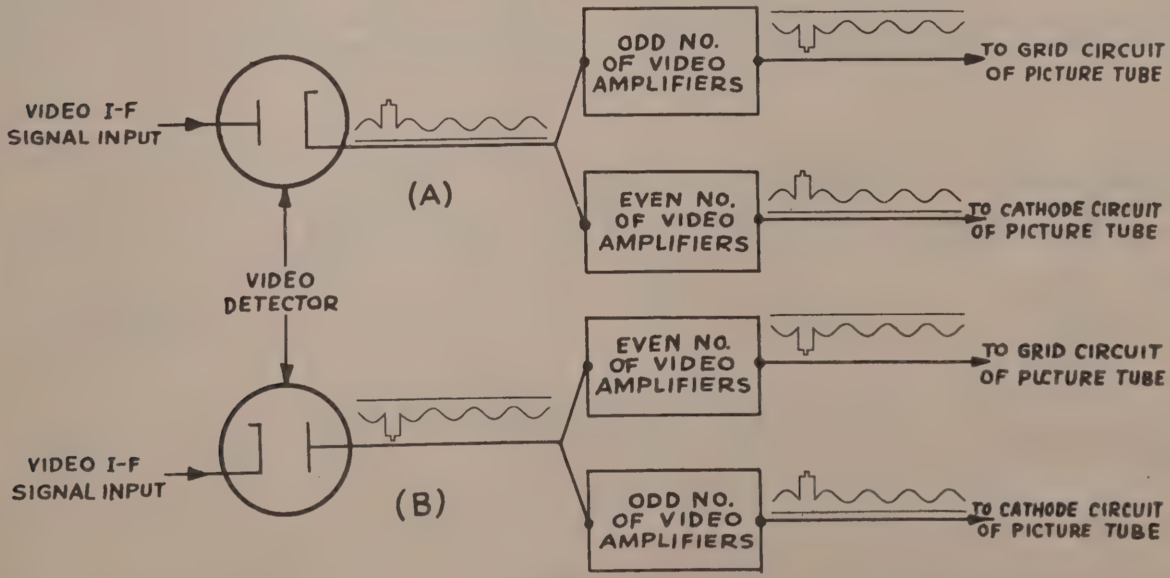

VIDEO I-F SIGNAL INPUT
(A)
ODD NO. OF VIDEO AMPLIFIERS
TO GRID CIRCUIT OF PICTURE TUBE
EVEN NO. OF VIDEO AMPLIFIERS
TO CATHODE CIRCUIT OF PICTURE TUBE
VIDEO DETECTOR
EVEN NO. OF VIDEO AMPLIFIERS
TO GRID CIRCUIT OF PICTURE TUBE
VIDEO I-F SIGNAL INPUT
(B)
ODD NO. OF VIDEO AMPLIFIERS
TO CATHODE CIRCUIT OF PICTURE TUBE

Fig. 6-27.—These various circuit arrangements are possible in the video section when the video signal output is fed to the picture tube from the plate circuit of the last video amplifier. The video modulating signal is assumed to be a sine wave.

region and that below the reference line in the negative or minus (—) region.

The diode will conduct only when the plate of the tube is effectively more positive than its cathode. Thus when the modulated i-f signal is fed to the plate of the diode Fig. 6-28 (A), *only* that part of the signal above the reference line will cause the diode to conduct. The negative part of the signal will make the plate negative with respect to the cathode and thus the diode will not conduct for this part of the input signal.

The output signal from the diode detector in Fig. 6-28 (A) is taken from the cathode circuit of the diode and is the positive half of the input signal. This output signal contains both the video signal and the high-frequency i-f component, and the latter signal is then removed by a filter network. The output of this filter network is the video modulating signal and it is then fed to the grid circuit of the first video amplifier.

When the video modulated i-f signal is fed to the cathode of the diode detector, only the negative part of signal will cause the diode to conduct. Without any signal applied to the diode, the plate and cathode are essentially at ground potential. When the potential on the cathode is made negative, the diode will conduct because the plate is effectively at a higher potential than the cathode. The output signal from the diode arrangement of Fig. 6-28 (B) is taken from the plate circuit of the tube and contains only the negative part of the input signal. This output signal contains the video signal and the high-frequency i-f component, and then the latter signal is removed by a filter network in similar fashion to that of Fig. 6-28 (A). The waveshape of the output signal from the filter network of Fig. 6-28 (B) is the video modulating signal, but it is of opposite phase to that of Fig. 6-28 (A). The two different types of detector arrangements of Fig. 6-28 provide either a positive or negative video signal output.

Amplifying vacuum tubes have the inherent quality of phase inversion between the input and output signals. If a signal voltage of zero degree phase is the input signal to the grid of a vacuum tube, the output signal voltage will be shifted in phase by 180°. Thus if the video signal output from the detector circuit of Fig. 6-28 (A) is fed to the grid circuit of a video amplifier, the output signal from this amplifier will be reversed in phase from that at its input. This is illustrated in the circuit of Fig. 6-29 (A). When the

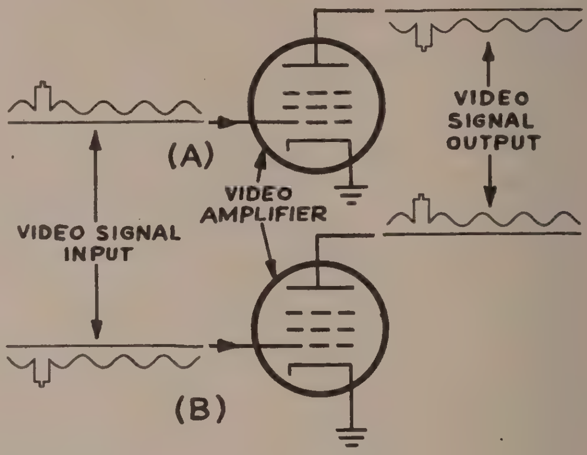


Fig. 6-29.—Because signals undergo a phase reversal of 180° between the input and output circuits of an amplifier tube, the video signals shown in Fig. 6-28 (A) and (B), when applied to the video amplifier grid, are reversed in phase in the amplifier output, as shown in (A) and (B) above.

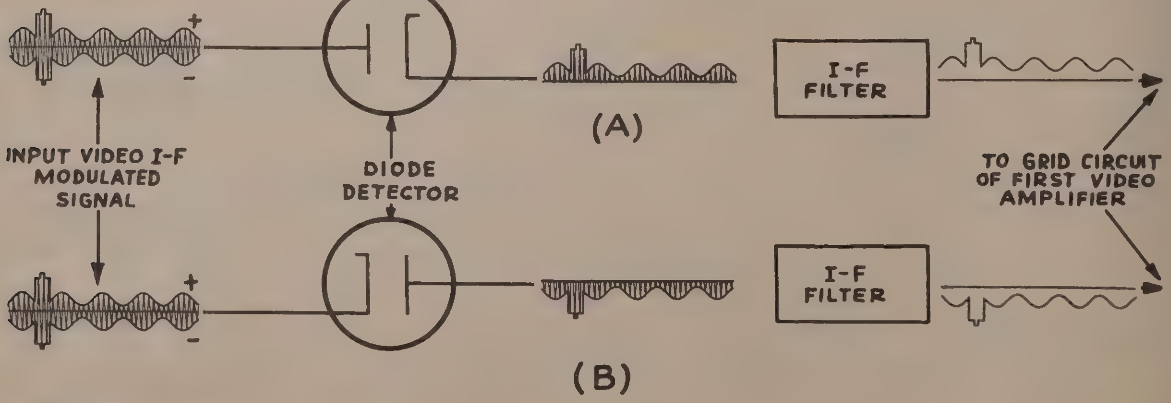

+
-
INPUT VIDEO I-F MODULATED SIGNAL
DIODE DETECTOR
(A)
I-F FILTER
TO GRID CIRCUIT OF FIRST VIDEO AMPLIFIER
+
-
I-F FILTER
(B)

Fig. 6-28.—How the phasing of the video modulated signal is controlled by the diode detector is shown here. In (A) the signal is fed to the plate of the diode rectifier which will conduct only when the plate is more positive than the cathode. That is, only that part of the signal above the reference line will cause the diode to conduct. When, as in (B), the signal is fed to the cathode of the detector, the reverse is true.

signal input to the amplifier is the output signal from the video detector circuit of Fig. 6-28 (B), the phase relationship between the input and output video signals of the amplifier is as shown in Fig. 6-29 (B).

If another amplifier were connected to the output of either circuit of Fig. 6-29 (where the coupling between these circuits does not introduce any appreciable phase shift), the added amplifier would merely reverse the polarity of the output signals illustrated in this drawing. The output signal from each second video amplifier would have the same waveshape and phase as the video signal input to their respective first video amplifiers

Detector Loading

Whatever the type of video detector used, the output loading circuit presents quite a design problem because of the frequencies involved. In a-m radio receivers the loading on the diode detector is usually quite high because the shunting capacities present in the circuit do not affect the audio-frequency response. In video detection the diode load has to be much lower because the shunting capacities present in the circuit affect the video-frequency response. The detected voltage is low because of this low diode load and this low voltage must be amplified to the level necessary for proper operation of the picture tube. Although the low impedance of the diode load causes the distortion introduced by this diode to increase somewhat, the added distortion introduced does not cause any bad visual defects in the reproduced picture since we are dealing with a video signal.

The video frequencies are as high as 4 mc. The total shunting capacities in the load circuit are high enough to offer a low impedance path to these high frequencies. Let us refer to the simplified diode detector circuit of Fig. 6-30. Resistor *R* represents the load resistance across the cathode of the diode. Capacitor *C* represents the total shunting capacitance in the circuit. This value of *C* is a controlling factor in the value of the load resistance. For example if *C* is 10 μμf, then at the 4-mc limit of the video signal the capacitive reactance will be approximately 4000 ohms. Thus if the video frequencies up to 4 mc must be passed with sufficient amplitude, the load resistance has to be much lower than 4000 ohms. If *R* is a higher value, the capacitive reactance of *C* will offer a lower impedance path to the high frequencies and will shunt them to ground. If *R* is lower, *C* will offer the higher impedance to the video frequencies and most of the signal voltage will appear across the resistor.

The resistance of *R* should be made appreciably lower than the reactance of *C* (1/5 is considered a good proportion) in order that the high video frequencies not be attenuated. For a capacitive reactance of 4000 ohms, resistor *R* should be about 800 ohms. The over-all diode load (*R* and *C* in parallel) should offer a fairly constant impedance at the video frequencies so that approximately the same output voltage is produced at these frequencies. Since the capacitance is the only factor in the parallel *RC* combination of Fig. 6-30 that varies with frequency, the effect of this capacitance must be diminished. To do this *R* should be decreased below the value of capacitive reactance.

Let us study two examples to see how the impedance of the *RC* combination can be kept fairly constant. The magnitude of the impedance in ohms (designated as *Z*) of a resistance and capacitance in parallel is given by:

$$Z = \frac{X_c R}{\sqrt{X_c^2 + R^2}} \text{ in ohms}$$

where X_c equals the capacitive reactance in ohms of *C* at the frequency in question and *R* the resistance in ohms.

Assuming that the capacitance is still about 10 μμf, then from the lowest video frequency of 30 cycles to the highest of 4 mc the reactance X_c will vary from about 530 megohms to approximately 4000 ohms. If the resistance of *R* is about 10,000 ohms, then the impedance *Z* for the highest video frequency will be:

$$Z = \frac{4000 \times 10000}{\sqrt{(4000)^2 + (10{,}000)^2}} = 3710 \text{ ohms}$$

At the lowest video frequency X_c is very large compared to *R*, and the impedance *Z* is considered to be equal to 10,000 ohms, or the resistance of *R*.

From this analysis, the impedance *Z* is seen to vary from 3710 ohms to 10,000 ohms. This means that the voltage output from the diode will vary appreciably over the complete range of video frequencies and the response level for the different video frequency will be anything but constant.

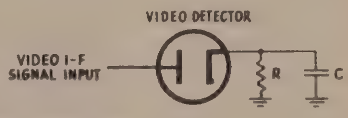


FIG. 6-30.—Simplified detector circuit wherein *R* is the load across the cathode and *C* represents the total shunting capacitance, the value of *C* being the controlling factor in the value of *R*.

If *R* is made smaller, the impedance will vary less over the range of video frequencies. If *R* is made equal to about 1000 ohms, then for the highest video frequency *Z* will be:

$$Z = \frac{1000 \times 4000}{\sqrt{(1000)^2 + (4000)^2}} = 970 \text{ ohms}$$

At the lowest video frequency X_c is very large compared to the 1000 ohms of R so that it has very little effect upon the magnitude of Z, and thus Z can be considered to be equal to 1000 ohms at the low video frequencies. Thus when the resistance R is lowered to 1000 ohms, the impedance of the load will vary by only 30 ohms (from 1000 to 970) instead of the former impedance variation of about 6300 ohms. The variation of 6300 ohms is about 60 percent of the maximum impedance of 10,000 ohms, and while the variation of 30 ohms is only 3 percent of the maximum impedance of 1000 ohms. From this comparison we can readily see that the lower the load resistor R (of Fig. 6-30), the more constant the over-all load impedance over the complete range of video frequencies and, hence, the more level the response.

The capacitance C in the previous analysis represents the total shunting capacitances. These shunting capacitances include the interelectrode and output capacitances of the diode detector, those due to the stray wiring of the circuit, and those of any circuit that is coupled to the output of the diode. The total value of these capacitances is quite high for the upper end of the video frequencies. Since they are inherent parts of the circuit, they are difficult to reduce; the diode load resistance, therefore, must be reduced to keep the output voltage from the detector constant over the complete range of video frequencies.

The amount the diode load resistance can be reduced is critical because the value of the total shunting capacitances C may run as high as 25 $\mu\mu f$. At 4 mc the capacitive reactance of 25 $\mu\mu f$ is approximately 1600 ohms, and the detector load resistance would have to be quite low for proper operation of the circuit. If this load resistance is too low, not enough signal voltage will exist in the output. Therefore we are faced with the problem of using the lowest possible value of load resistance to provide adequate signal output and yet to keep this output voltage as nearly constant as possible over the complete range of video frequencies.

Since insufficient signal output exists when the value of the detector load resistor is too low, this resistor may have to be somewhat higher than anticipated. In practice this load resistor generally varies from about 1500 ohms to 5000 ohms. There is appreciable attenuation of the higher video frequencies with these values of resistances. This is undesirable, especially in those television receivers that employ large screens because discrimination against the high video frequencies will be noticeable on the reproduced picture. In order to make the level of all the video frequencies as constant as possible, many diode detector circuits use some form of *peaking circuit* in the output of the detector. These peaking circuits essentially consist of coils placed in the output of the detector circuit (part of the coupling system to the first video amplifier) and their values are so chosen that they will resonate with the shunting capacitances. (In some cases a shunt capacitor is added to develop the desired resonance.) Resonance occurs at the high video frequency end and tends to peak the response curve there. Resistors are usually shunted across these peaking coils to reduce the height of the peak to the average level of the video response curve. A complete detailed mathematical analysis of peaking circuits is included in the next chapter.

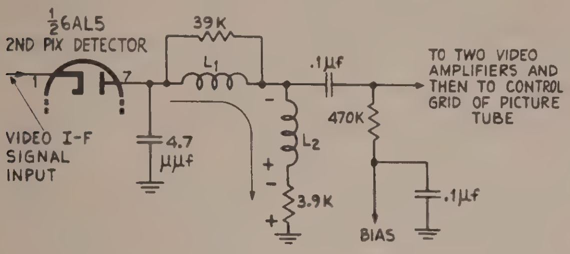
In the following sections we will analyze a number of different video detector circuits used in today's receivers with regard to the type of diode connections, number of video amplifiers, grid or cathode signal injection to the picture tube, loading on the diode, and the coupling circuit from the detector to the first video amplifier. In all of the circuits to be discussed, the video signal fed to the picture tube is secured from the plate circuit of the last video amplifier.

Garod Model 3912 — TVFMP

In the video detector circuit of the Garod television receiver model 3912 — TVFMP, one half of a 6AL5 duo-diode tube is used as the detector and the input video i-f signal is fed to the cathode (pin 1) of the tube. This may be seen from the schematic of Fig. 6-31. The loading on the plate of this tube consists of a 4.7 $\mu\mu f$ capacitor, two peaking coils L_1 and L_2, a damping resistor of 39,000 ohms across L_1 and a 3900-ohm plate load resistor. The remainder of the coupling network consists of a 0.1-$\mu\mu f$ coupling capacitor and the 470K grid-leak resistor of the following video amplifier stage.

The two coils in conjunction with the fixed 4.7-$\mu\mu f$ capacitor and the inherent shunting capacitances of the circuit are so chosen that they will resonate at the high-frequency end of the video signal, thereby helping produce a level video response curve. The 39,000-ohm resistor across the series peaking coil L_1 is employed to reduce or damp the peak caused by the resonance of coil L_1 and its associated capacitance to further level the response curve.

The output from the detector system is fed into two video amplifiers and then into the grid circuit of the picture tube. By comparing the video detector and amplifier system of Fig. 6-31 with that of Fig. 6-27, it will be seen that the correct polarity of the signal input to the grid of the picture tube is maintained. This polarity, as established, should be negative. If an even number of video amplifiers is employed, then the input video signal to the control grid circuit of the first amplifier should likewise be negative. This negative polarity is obtained by feeding the video i-f signal input to the cathode of the diode detector. This was analyzed in connection with Fig. 6-28 (B) but it can also be seen from Fig. 6-31. In this latter circuit, the diode current flow is from cathode to plate and therefore takes the path indicated by the arrow. The polarities that appear across coil L_2 and the 3900-ohm load resistor are as indicated in Fig. 6-31, and thus a negative polarity will exist at the junction of L_1 and L_2 with respect to ground.

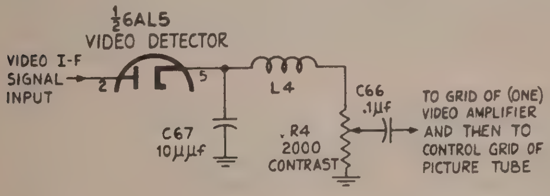


After Garod Radio Corp.

FIG. 6-31.—The video detector circuit of the Garod model 3912-TVFMP wherein the video i-f signal is fed to the cathode of the 6AL5 detector.

Belmont Model 21A21

The video detector circuit for the Belmont model 21A21 is illustrated in Fig. 6-32. The video i-f signal is fed to the plate of the diode detector which uses one half of a 6AL5 duo-diode, the other half being used for automatic gain control (agc). The final video amplified signal is fed to the control grid of the picture tube. Due to this arrangement an odd number of video amplifiers has to be employed to obtain the necessary polarity of the video signal input to the grid of the picture tube. In this receiver one video amplifier is employed. This system is representative of the block diagram circuit at the top portion of Fig. 6-27 (A).



After Belmont Radio Co.

FIG. 6-32.—The video detector circuit of the Belmont model 21A21, wherein the video i-f signal is fed to a plate of the 6AL5 duo-diode.

The interesting part of this circuit is the loading network in the cathode output of the tube. Coil L_4 is a series peaking coil and in conjunction with the 10-*μμf* fixed capacitor *C*67 and the other shunting capacitances in the circuit it is used to *peak* the high-frequency end of the curve. The loading resistor in this circuit is a 2000-ohm potentiometer *R*4, across which most of the output voltage appears. The variable arm of this potentiometer is connected to the grid circuit of the following video amplifier tube through a 0.1-*μμf* coupling capacitor *C*66. Thus the variable arm controls the amount of voltage input to the video amplifier and, hence, the video signal input to the picture tube and thus functions as a *contrast control*.

General Electric Model 802

In the GE model 802, the video signal input is fed to the cathode circuit of the picture tube. An odd number (one) of video amplifiers is employed so that the video i-f signal input to the diode detector must be fed to the cathode of the circuit in order to have the proper polarity of video signal to the picture tube. This was shown in the lower block diagram circuit of Fig. 6-27 (B).

The video detector circuit for this receiver is illustrated in Fig. 6-33. The detector utilizes one half of a 6H6 duo-diode tube. The load resistor *R*18 in the plate circuit of this detector is quite low, being 1500 ohms. This resistor is shunted with a fixed capacitor, *C*33, of 10 *μμf* in conjunction with the inherent shunting capacitances that already exist across the resistor. Series coil *L*10 is employed to peak the video response curve at the necessary video high-frequency end. The interesting thing about this circuit is that this peaking coil, or coupling coil as it is sometimes called, is inserted after the load resistor.

Peaking coils are placed in different parts of the coupling circuit between the output of the video detector and the video amplifier (as well as between video amplifiers) to separate the inherent shunting capacitances that exist in the circuit. The reason for

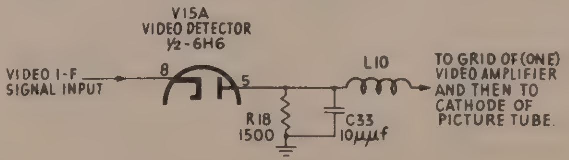

V15A
VIDEO DETECTOR
½-6H6
VIDEO I-F SIGNAL INPUT
8
5
L10
TO GRID OF (ONE) VIDEO AMPLIFIER AND THEN TO CATHODE OF PICTURE TUBE.
R18
1500
C33
10µµf

FIG. 6-33.—The video detector circuit of the GE model 802 feeds the video i-f signal through one video amplifier to the cathode of the picture tube.

the variation in placement is that the shunting capacitances in the circuit are divided by the series peaking coil, and the amount of shunting capacitance on either side of the coil varies in different circuits. The placement of the coil depends upon obtaining the correct capacitance needed to form the resonant peaking circuit.

In all of the detector circuits discussed so far, the fixed value of capacitance inserted in the diode load serves as a filter for the video i.f. The output signal from the detector is a video-modulated i-f signal that has gone through the process of rectification. It contains the video modulating frequencies — up to 4 mc (if the over-all video i-f response is designed to pass the high frequencies) and also the carrier frequency which has been converted to an i.f. The video i-f carrier can be considered in round numbers to be about 25 mc. Only the video modulating frequencies are to be passed and not the i.f. The fixed capacitance in the diode load is of a high enough value to present a low impedance path to ground at the video i.f. and therefore filter it to ground.

Consolidated Television Model 2315

In the Consolidated Television Corporation model 2315 a 1N34 germanium *crystal* is employed as a video diode detector. The diagram for this circuit appears in Fig. 6-34. The terminal of the crystal marked with a negative (—) sign represents the cathode and that with the positive (+) sign represents the plate of the detector. Two video amplifiers are employed in this receiver and the video signal from these amplifiers is fed to the grid of the picture tube. With this arrangement, for the proper polarity of video signal input to the grid of the picture tube, the video i-f signal must be fed to the cathode, or negative terminal, of the crystal as indicated in Fig. 6-34.

A shunt peaking coil *PL*-1 and a series peaking coil *PL*-2 are used in the output of this circuit to increase the response at the high frequencies. The load resistor is *R*14, 3300 ohms. The use of a 1N34 germanium crystal as a diode detector reduces the over-all inherent shunting capacitances in the circuit. There is no heater to cathode, nor cathode to plate interelectrode capacitance as when a regular diode is used. There are some receivers on the market that employ a 1N34 crystal as a video detector and do not use any peaking circuits in the output because of the reduced shunting capacitances in the circuit. A higher value load resistor can then be used in the detector stage.

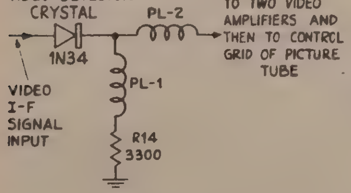

VIDEO DETECTOR
CRYSTAL
PL-2
1N34
TO TWO VIDEO AMPLIFIERS AND THEN TO CONTROL GRID OF PICTURE TUBE
VIDEO I-F SIGNAL INPUT
PL-1
R14
3300

FIG. 6-34.—In the Consolidated Television model 2315 a 1N34 germanium crystal is used as a video diode detector.

Television Assembly Model F1-101

The video detector circuits discussed so far have employed one half of a duo-diode 6H6 or 6AL5 tube or a crystal rectifier as the detector. The video detector circuit of the Television Assembly model F1-101 uses a detector system different from anything yet discussed. A 6J6 *duo-triode* is employed in a video detector system which utilizes both triode sections of the tube.

The circuit for this unique arrangement appears in Fig. 6-35. Although not shown, the receiver has two video amplifiers, and the output from the plate circuit of the last amplifier is fed to the grid circuit of the picture tube. In order to obtain the correct polarity of the video signal input to the grid of the picture tube with this system, the video i-f signal should be fed to the cathode circuit of a diode detector. This was illustrated in the block diagram of the top half of Fig. 6-27 (B). At a first glance at the detector circuit of Fig. 6-35, it may be difficult to see what type of detection is employed. However, if this circuit is studied carefully, it will be seen to be a diode rectifier which is driven by a cathode-follower stage. Let us now analyze this circuit in detail.

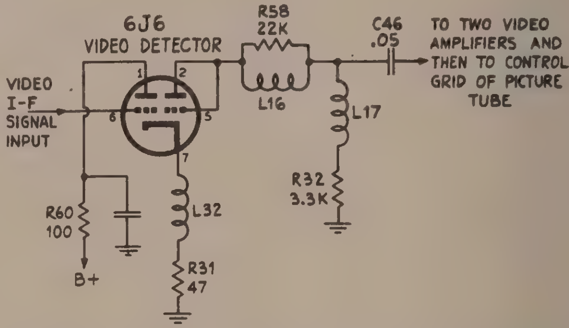


FIG. 6-35.—Unlike any of the models previously illustrated, the Television Assembly model F1-101 employs a duo-triode 6J6 tube as a video detector. The circuit is a diode rectifier driven by a cathode-follower stage, with the first triode section acting as the cathode-follower driver and the second as the diode detector.

The first triode section of this tube is used as the cathode-follower driver. The video i-f signal output from the last video i-f stage is fed to the grid, pin 6, of the 6J6 detector tube. The series circuit in the cathode consists of a 47-ohm resistor, *R*31, and coil *L*32. This network is unbypassed and the video i-f signal appears across this cathode load. The other triode section of the tube has its plate and grid tied together and in conjunction with its cathode will act as a diode. This section of the tube acts as the video diode detector.

Since the cathode of the 6J6 tube is common to both triode sections, it represents a ready source from which the diode section can obtain its video i-f signal. The cathode-follower system is one method that results in a fairly high video i-f signal input to the detector so that, with the usual low load impedance on the detector, an adequate amount of rectified voltage output is still obtained.

The video i-f voltage drop across the cathode circuit is the input signal to the detector. The output impedance of the cathode follower is low compared to the series impedance of the detector section and the load on the plate of the detector. This output impedance of the cathode follower (in this case considered as the source impedance, since the input or source voltage appears across this circuit) and that of the detector plus the plate load present a voltage dividing network to the detected signal. This means that after detection most of the detected video signal appears across the load.

The load of the detector essentially consists of the series peaking circuit of *L*16 in parallel with *R*58, the shunt peaking coil *L*17, and the plate load resistor *R*32. The output video detected signal is coupled to the following video amplifiers through the 0.05-μf capacitor *C*46.

Hallicrafters Model T-54

In the video channel section of the Hallicrafters model T-54 is employed a unique video i-f, detector, and amplifier system. Because the video diode detector is tied up with the operation of the entire system, the discussion of this receiver is included in this chapter. One half of a 6H6 duo-diode is used as a *diode detector* and also as a *diode mixer* and source for *automatic gain control* (agc). The output from the last video amplifier stage of this unit is not taken from the plate of the tube as was the case in all the other systems discussed but rather from the cathode. All these features will now be discussed in detail.

A schematic diagram of the video detector and amplifier unit is illustrated in Fig. 6-36. Although not shown here, all the i-f amplifiers employed in the video section can be considered as composite amplifiers. Ordinarily, separate i-f amplifiers are employed for video and sound to prevent interaction between the two signals. In this receiver, however, the interaction is put to use; the two i-f carriers are fed into the video diode detector circuit and beat together to produce a new i-f signal.

The composite sound and video i-f signal is fed to the cathode of the diode as seen in Fig. 6-36. The bandwidth of the video i-f system is narrow and some high-frequency video signals are attenuated. However, the seven-inch picture tube that is employed still has adequate definition of the picture even with the loss of some of these high video frequencies. The accompanying sound i.f., being on the high-frequency end of the video i-f response curve, is also attenuated but enough accompanying sound i-f signal does pass

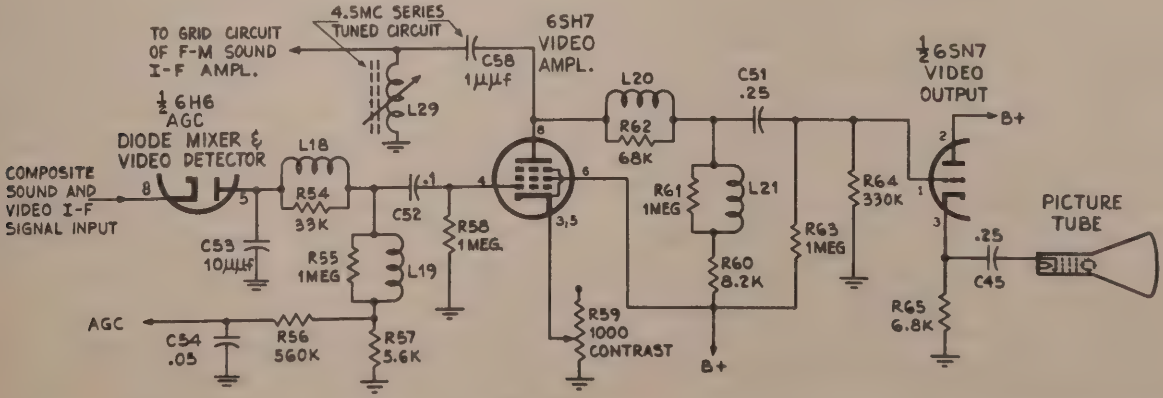


Fig. 6-36.—The video detector and amplifier unit of the Hallicrafters model T-54 makes use of the interaction of the video and sound signals, and the composite signal is fed to a 6H6 duo-diode tube which functions as diode detector, diode mixer, and source for agc.

through these circuits and enters the video detector with the video i-f signal.

In this receiver the video i-f carrier is 26.25 mc and the sound i-f carrier is 21.75 mc, the difference being 4.5 mc, the standard separation between the video and sound carriers. As mentioned, these two i-f signals are mixed together in the diode detector and the resulting beat signal is 4.5 mc. Besides the diode acting as a mixer, it also acts as a rectifier to the a-m video i-f signal. Thus detection of the video signal also occurs in this tube. The 4.5-mc beat signal together with the video detected signal is coupled to the following 6SH7 video amplifier stage, as seen in Fig. 6-36.

The coupling circuit is an *RC* stage to which series and shunt peaking circuits have been added to peak the high frequencies. The series peaking circuit consists of coil *L*18 in parallel with *R*54 and the shunt peaking circuit consists of *L*19 in parallel with *R*55. The 5600-ohm diode load resistor, *R*57, is in series with the shunt peaking circuit, and both together offer a fairly high impedance load to the diode so that appreciable voltage appears between the junction point of *L*18, *L*19, and ground. The composite signal, the 4.5-mc beat signal and the detected video, is fed to the grid circuit of the 6SH7 video amplifier through the 0.1-μf coupling capacitor *C*52.

Attached across *R*57, the diode plate-load resistor, is a resistance capacitance network, *R*56-*C*54, which acts as a filter to the alternating video signal. This filter is similar to the avc filters used in a-m radio receivers in which the audio is filtered out and a negative d-c signal remains which is used for avc. Thus in Fig. 6-36, *R*56 and *C*54 filter out the alternating video variations and a negative d-c signal remains which is used for agc on the first and second video i-f amplifiers. (It should be remembered that the composite input signal to the cathode of the detector tube has a d-c component and therefore agc is possible.)

Let us now return to the 6SH7 first video amplifier. The input signal, the 4.5-mc beat signal and the video modulating signal, is amplified by the tube. The cathode resistor, *R*59, of this tube is an unbypassed 1000-ohm potentiometer, and thus degeneration occurs in this tube. By varying the value of this resistance, the amount of degeneration and bias of the tube is changed and the gain of the tube can be controlled. This potentiometer therefore acts as the contrast control of the receiver. The approximate gain of the tube is about 20 to 25 times.

The signal resulting from mixing the 26.25-mc a-m signal (the video i-f carrier) and the 21.75-mc f-m signal is a 4.5-mc beat signal which varies in frequency and amplitude. Although the beat signal voltage is small compared to the video signal, it will have sufficient strength after amplification by the 6SH7 video amplifier to be used as the *sound i-f signal*. Directly at the plate circuit of this video amplifier is a series tuned *LC* circuit, consisting of capacitor *C*58 and variable inductance *L*29, which is tuned to 4.5 mc. This series tuned circuit, therefore, acts as a low impedance path to ground for the 4.5-mc beat signal. Since this is a series tuned circuit, maximum current will flow at resonance and a maximum voltage drop will appear across *each* element in the circuit. The 4.5-mc signal is tapped off across *L*29 and fed to the grid circuit of the only sound i-f amplifier in the circuit. From this sound i-f amplifier, the signal is fed to a ratio detector circuit. This circuit responds only to f-m and not to a-m variations in the input signal. The f-m variations of this signal are then detected and applied to the audio amplifiers of the receiver. A television sound system of the type just discussed, where both the sound and video i-f signal output of the frequency conversion system beat together to form a new sound i-f signal, is commonly known as an *intercarrier* sound system, since it depends for its action upon intermodulation of the sound and video (i-f) carriers.

The *C*58-*L*29 series-tuned 4.5-mc circuit prevents the beat signal from interfering with the video signal reproduction. Thus only the video signal output from the plate circuit of the first video amplifier is coupled to the video output tube. As seen in Fig. 6-36, the series and parallel peaking networks, *L*20-*R*62 and *L*21-*R*61 respectively, also appear in the plate circuit of this tube. The video signal is finally coupled to the video output tube, which is one triode section of a 6SN7. This output tube functions as a cathode follower where the cathode load is just the 6800-ohm resistor *R*65. The output from this cathode-follower circuit is applied to the cathode of the picture tube through the 0.25-μf coupling capacitor *C*45.

Compensating networks are not needed because the video output signal is taken from the cathode and not the plate circuit of the video output stage. The output tube does not contribute any gain when acting as a cathode follower, but the preceding video amplifier provides enough gain for proper video signal reproduction.

The type of coupling system to the picture tube is completely different from any of the circuits previously analyzed, so it does not fall into any of the categories of Fig. 6-27. However, for proper operation of the picture tube, the polarity of the video signal input to its cathode should be positive with respect to its grid. The proper polarity is obtained from this circuit but this may not be readily evident. Let us analyze the complete circuit of Fig. 6-36 to see how this occurs.

The input video modulated i-f signal to the diode detector is fed to the cathode of the tube. Only the negative part of this signal will cause diode current to flow because it effectively makes the plate more positive than the cathode. The positive part of the signal is thus clipped, and the negative video signal is then coupled from the diode plate circuit to the grid of the 6SH7 video amplifier with negligible phase change. This negative video signal is reversed in phase by 180° due to the inherent phase reversing qualities of the tube and appears in the plate circuit as an amplified, but positive polarity, video signal. This signal is next coupled to the grid circuit of the cathode-follower video output tube. Negligible phase change occurs between the grid and plate circuit of this tube, so that the positive polarity that exists for the video signal at its grid also exists at its cathode. Since the cathode of the picture tube is connected to the cathode of the video output tube, it receives a positive video signal which is necessary for the correct operation of the picture tube.

CHAPTER 7

VIDEO AMPLIFIERS AND D-C RESTORERS

By Henry Chanes

The signal at the output of the video second detector contains all the information necessary to form the picture on the face of the picture tube. The sync signals will be removed from this signal and used to sync the sweep circuits; this is discussed in chapter 8. We are now mainly interested in the blanking signals and the picture components of the signal at the output of the video second detector. The blanking signals are used to drive the cathode-ray-tube grid to cutoff during both horizontal and vertical retrace. The spot on the screen is thus extinguished during the retrace time to prevent interference with the picture. The picture components of the signal determine the brightness of each small picture element. These elements, in turn, make up the complete picture. The signal available at the output of the video second detector is usually only about 1 volt, peak-to-peak. This is insufficient to drive the picture tube properly, and therefore this signal has to be amplified. The function of the video amplifier is similar to that of the audio amplifier in a receiver, but its construction is complicated by stricter requirements. Let us examine these requirements.

Requirements of a Video Amplifier

The video amplifier must amplify equally well all frequencies from as low as 30 cycles to as high as 4 mc. A good audio amplifier has to cover a range of only about 70 to 10,000 cycles. Fig. 7-1 illustrates the different bandwidths required by the two amplifiers. The dashed line is the frequency response required of the video amplifier, and the solid line represents the response of a good audio amplifier.

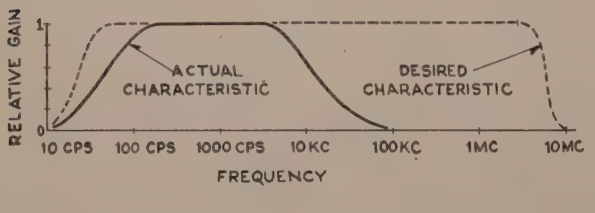


Fig. 7-1.—The solid line represents the bandwidth of an audio amplifier. The dotted line represents the desired bandwidth of a video amplifier.

In an ordinary audio amplifier, the amount of phase shift of a signal going through the amplifier is not an important consideration, as the ear is not capable of detecting phase distortion. In a video amplifier this is a very important consideration. Let us consider phase shift in terms of time delay. The time that the signal is delayed in going through the amplifier is very small, being of the order of microseconds (millionths of a second), when compared to our usual concept of time. However small this time delay may seem, it is very important. Consider a 12-inch picture tube with a picture width of approximately 9 inches. The spot moves across the screen 15,750 times a second, or at the rate of 142,000 inches per second. If a particular frequency in the video signal that represented a light and dark pattern were delayed by only 1 microsecond, the pattern would be displaced on the screen by about 0.14 inch. If every frequency were delayed the same amount of time, the delay would not be noticeable, as the entire picture would be moved over a little. However, if different frequencies were delayed by different times, the distortion in the picture would be immediately apparent. The relation between the phase shift and the time delay is given by

$$T = \frac{\theta}{360 f} \qquad \text{Eq. 7-1}$$

where T = time delay in seconds
θ = phase shift in degrees
f = frequency in cycles

From this formula it can be seen that to keep the time delay constant for all frequencies, the phase shift must increase in proportion to the frequency. This is known as linear phase shift and is illustrated in Fig. 7-2.

The gain necessary in a video amplifier depends on the output of the video second detector and the amount of signal necessary to drive the grid of the picture tube properly. This gain varies from about 25 for a 7-inch tube to about 50 for a 10- or 12-

inch picture tube. The number of stages necessary to achieve this gain is usually one or two.

The other requirements for the video amplifiers are those associated with any amplifier. They include minimum harmonic distortion, a satisfactory signal-to-noise ratio, and decoupling when necessary to prevent oscillation. These offer no special problem, except where complicated by the necessity of a large bandwidth and linear phase shift.

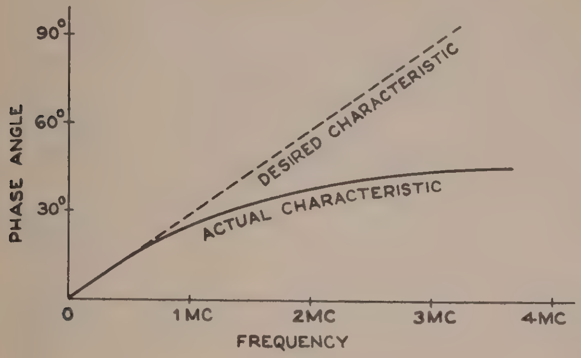


Fig. 7-2.—The dotted line represents the desired linear phase shift and the solid line represents the actual phase shift as it varies with frequency.

Basic Circuit for Video Amplifier

The basic circuit for a video amplifier is an *RC* coupled amplifier as shown in Fig. 7-3. This circuit has no compensation for either the high- or low-frequency response. The gain will fall off at both the high- and low-frequency ends, and the phase shift of the signal will not be proportional to the frequency. The amplitude and phase response characteristics of a typical amplifier of this type are shown in Figs. 7-1 and 7-2. The solid line represents the actual response, and the dashed line represents the response desired in the video amplifier. Examining the amplitude response, it can be seen that the response will have to be increased or "peaked" for both the low- and the high-frequency ends of the frequency range. Let us briefly discuss the reasons for the response dropping off in this basic amplifier.

Referring to Fig. 7-3 and considering the low frequencies, we can see that the coupling capacitor C_b is in series with the grid resistor R_c of the following tube. The output signal from the first tube is applied across this series combination. As the frequency becomes lower, the reactance of the capacitor C_b becomes higher, and less of the output signal appears across resistor R_c to be applied to the next tube. There is also a phase shift due to this resistor-capacitor combination.

At the high frequencies the effect of C_t becomes appreciable. C_t represents the total shunt capacity due to the output capacity of the first tube, the input capacity of the second tube, and all the stray capacity due to the wiring. At the high frequencies the reactance of this capacitance becomes small enough to lower the impedance of the plate load on the tube. The plate load, instead of being a pure resistance, will then be a resistance in parallel with a capacitive reactance. This smaller load impedance will cause the output voltage to drop and the capacitive component of the load will cause a phase shift in the output signal.

Because of the deficiencies mentioned above, it is evident that the ordinary *RC* coupled amplifier is not good enough for use as a video amplifier without some form of compensation to offset the falling off in response at the low and high frequencies and to give a linear phase shift to prevent phase distortion. We will now consider the different methods used for low- and high-frequency compensation.

Low-Frequency Compensation

We have seen how the combination of coupling capacitor C_b and grid-leak resistor R_c reduced the low-frequency response of the video amplifier. These effects of attenuating the low frequencies must be compensated for in video amplifiers. Video amplifiers employ pentode tubes of such design that they have relatively high values of mutual conductance, g_m, and exceptionally small values of interelectrode capacitances. The reasons for these design features will soon be evident. The background of the reproduced picture seen on the screen of a television receiver is due to the very low frequencies of the video signal received. If these low frequencies were attenuated, the picture would not be in good *contrast* to its background and this is undesirable. Before going into a circuit analysis of low-frequency compensation, it should be stated that it is difficult

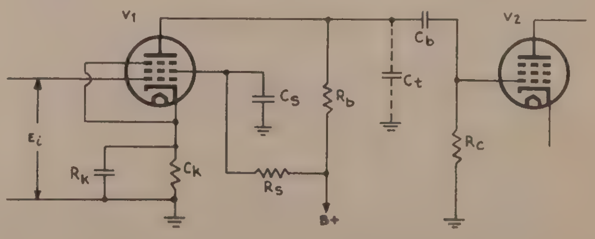


Fig. 7-3.—Typical *RC* coupled amplifier.

to test video circuits for proper low-frequency response with the input and output measuring instruments that are generally used for testing the response of ordinary *RC* couplings. In checking the degree of attenuation at the low frequencies, it is well to use a low-frequency square-wave input and to observe the distortion of the output signal waveform on an oscilloscope. Such square-wave analysis should cover the range from 20 cycles to 1000 cycles. When the amplifier is properly compensated, the square-wave output resembles the square-wave input very closely. (On an oscilloscope distortions as low as 5% can be recognized. This is considered accurate for design purposes.)

When an *RC* coupled amplifier is compensated at the low-frequency end, another resistance R'_b is placed in series with the plate-load resistor R_b and a capacitor is placed across the former resistor. The reasons for the use of these extra elements and the typical values chosen will be discussed. In Fig. 7-4 is shown a typical low-frequency compensated *RC* coupled amplifier in which $R'_bC'_b$ is added to better the low-frequency response. This addition of $R'_bC'_b$ appears to be nothing more than a decoupling network. In reality it is a decoupling network with the screen $B+$ lead connected directly to $B+$ instead of between R_b and R'_b. Besides its immediate use of raising the low-frequency response, it decouples any undesired feedback voltage due to the common impedance of the plate supply.

Since the tube used is a pentode, we can make some approximations for simplifying the analysis of this circuit. The grid-leak resistor R_c is much larger than the plate-load resistor R_b ($R_c >> R_b$) and the plate resistance of the tube R_p is much greater than R_b ($R_p >> R_b$); therefore all that is necessary is to take into consideration the essential effect of the combination R_cC_b. At low frequencies R_cC_b causes a drop in signal potential from the plate of tube V_1 to the grid of tube V_2 and also phase shift between the two tubes. The coupling capacitor C_b causes the shift. Why, then, is the value of C_b not made large enough so that it will offer negligible resistance at low frequencies? The value of C_b could be increased and this would improve the low-frequency response, but since we are dealing with frequencies up to 4 mc, the larger the value of C_b, the greater will be the shunt capacitance to ground that it introduces. These shunt capacitances, although extremely small, are very detrimental at the high frequencies. Low-frequency compensation is used to avoid the introduction of these shunt capacitances. The insertion of $R'_bC'_b$ compensates for the loss in gain at the low frequencies and also introduces a phase shift to compensate for the undersired phase shift introduced by R_cC_b.

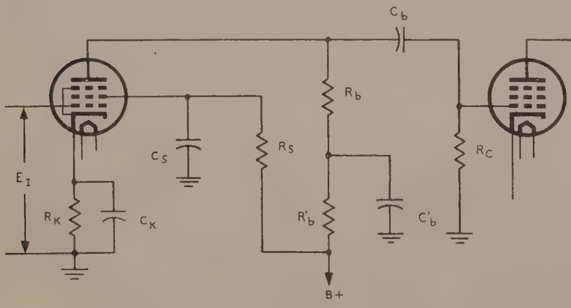


FIG. 7-4.—Typical low-frequency compensated *RC* coupled amplifier. The decoupling network $R'_bC'_b$ has been added to the circuit of Fig. 7-3 to better the low-frequency response.

Since the combination of R_cC_b at low frequencies reduces the gain, the combination of $R'_bC'_b$ at these same frequencies is inserted to offer a higher load impedance to tube V_1, and increase the gain. Increasing the load impedance will increase the effective voltage across the load and bring the gain back to normal. Again referring to Fig. 7-4, the operation is as follows: the voltage across the grid-leak resistor decreases (thereby decreasing the input to the grid of tube V_2) because the reactance of C_b increases as the frequencies are decreased. The reactance of capacitor C'_b increases at the same time. As a result, the parallel combination of $R'_bC'_b$ is seen to be an impedance that is added to the plate-load resistor R_b and effectively increases the total load impedance on tube V_1 as the frequencies are decreased.

Since this *RC* coupled amplifier stage uses a pentode, then $R_p >> R_b$ and the gain at low frequencies in the case of this compensated stage is expressed as follows:

$$A_{low} = A_{int} \frac{1}{\sqrt{1 + \left(\frac{X_{cb}}{Z_{load}}\right)^2}} \qquad \text{Eq. 7-2}$$

The letter A is used to designate the gain of a stage, and X is the expression for reactance and Z that for impedance as usual. Therefore:

A_{low} = the gain at low frequencies
A_{int} = the gain at intermediate frequencies
X_{cb} = the reactance of the capacitor C_b
Z_{load} = the total effective load impedance, including R_b, R'_b and C'_b

From this equation it may be seen that the effect of the original plate-load resistor R_b has been changed to Z_{load}. Since the plate-load impedance has been increased by the $R'_bC'_b$ combination, the gain at low frequencies is also increased. As the frequency decreases, the ratio of X_{cb}/Z_{load} also decreases, making Eq. 7-2 approach the gain at intermediate frequencies, A_{int}. It would be ideal if resistor R'_b could be increased to an infinite amount but the need for an appreciable amount of plate voltage on the tube makes this impossible. If this could be achieved, the gain would be equal to that of the tube itself and no loss would be encountered; the gain would be maximum and flat throughout the response. From *experimentation* it has been found that for the proper response the time contants of $R'_bC'_b$ and R_cC_b should be equal to each other. The voltage loss due to C_b will then be compensated (in direct proportion) by the voltage gain due to the total load impedance, in such a way that the voltage across the grid-leak resistor R_c will remain constant. In many instances the combination of $R_bC'_b$ is chosen first and the coupling capacitor C_b is made as large as is desired, then the grid-leak resistor R_c is determined. For instance, assuming that C'_b is chosen to be 30 μf, C_b chosen as 0.1 μf, and the time constant of R_cC_b to be 0.04 second (or 40,000 microseconds) it is desired to know what the values of R_b, R'_b and R_c are, at the lowest frequency of operation, which in this case is 30 cycles. Therefore:

$$R_cC_b = R_bC'_b = 40{,}000 \text{ microseconds,}$$

and since $C'_b = 30\ \mu f$

$$R_b = \frac{40{,}000}{C'_b} = \frac{40{,}000}{30} = 1333 \text{ ohms}$$

and since $C_b = 0.1\ \mu f$ then:

$$R_c = \frac{40{,}000}{0.1} = 400{,}000 \text{ ohms}$$

R'_b should be chosen to be about 20 times greater that the reactance of C'_b at the lowest frequency to be passed (30 cycles in this case) and yet not drop too much of the supply voltage. Therefore:

$$R'_b = 20\,X_{C'b} = 20 \times \frac{1}{2\pi f C'_b}$$

$$= \frac{10}{(\pi)\ (30)\ (30)\ (10^{-6})} = 3540 \text{ ohms}$$

We therefore find that:

R_b——should be about 1300 ohms
R_c——should be about 400,000 ohms
R'_b——should be about 3500 ohms

The equation for the gain at low frequencies of an uncompensated RC coupled amplifier is:

$$A_{low} = A_{int}\frac{1}{\sqrt{1+\left(\frac{X_{cb}}{R_c}\right)^2}} \qquad \text{Eq. 7-3}$$

If this equation is studied, it will be found that the phase shift, due to the R_cC_b combination, at the low frequencies is:

$$\theta = \tan^{-1}\frac{X_{cb}}{R_c} \qquad \text{Eq. 7-4}$$

where θ is in degrees, and the expression $\tan^{-1}$ means the angle whose tangent is the quantity following the expression. It can be seen that the coupling capacitor C_b causes a lead in phase. Inserting the $C'_bR'_b$ network into the load tends to cancel this phase lead by introducing a phase shift in a lagging direction.

Besides the grid-leak resistor and coupling capacitor combination, there are two other resistance-capacitance combinations that cause a falling off in gain and a leading phase shift at the low frequencies. First, there is the effect of the R_KC_K bias impedance. The best way to eliminate this effect would be to eliminate R_KC_K and apply bias another way. This is usually not desired as cathode bias is preferable in many cases, so it has been determined experimentally that this undesired effect can be compensated for by making the time constant of R_KC_K equal to the time constant of $R'_bC'_b$. Mathematically we have:

$$R_KC_K = R'_bC'_b \qquad \text{Eq. 7-5}$$

Secondly, there is the effect of the screen grid circuit. A pentode tube possesses a dynamic screen grid resistance as well as a dynamic plate resistance. This screen grid resistance R_{SG} (so called to differentiate it from the screen dropping resistor R_s) in combination with the bypass capacitor C_s also causes a falling off in response and a phase shift (leading) at the low frequencies. The effect of this combination is similar to that of the R_cC_b combination and experimentally it has been found that for the phase shift not to exceed 2° at 60 cycles, the time constant of $R_{SG}C_s$ should be greater than 0.076 second. Since it is difficult to vary such an element as R_{SG}, and because C_s is the only variable parameter, little compensation can be accomplished by a single stage; better compensation is obtained by the use of several stages of amplification.

A very important thing to note is that in all these compensating problems $R'_bC'_b$ plays the important role. Although $R'_bC'_b$ can be used to compensate for any one of the above cases, only one case can be compensated for in a stage such as shown in

Fig. 7-4. A usual practice in cathode biased amplifiers is to make R_c very large (this is usually possible) and, therefore, the time constant of R_cC_b will be large and the needed compensation is shifted to the R_KC_K combination (instead of the $R_cC_b=R_bC'_b$ combination).

In Fig. 7-5 are several curves showing the effect of the insertion of the $R'_bC'_b$ network on the low-frequency response for different values of $R'_bC'_b$.

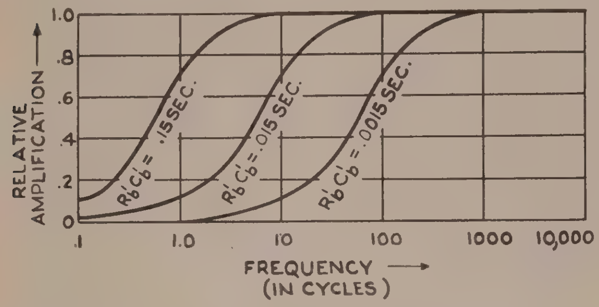


FIG. 7-5.—The effect of inserting different values of the $R'_bC'_b$ network is shown. As the time constant of the decoupling network is increased, the low-frequency response increases.

From these curves it is seen that as the time constant of the $R'_bC'_b$ combination is increased, the low-frequency response increases to frequencies as low as a few cycles per second. However, there are practical limitations to the values of R'_b and C'_b that can be chosen.

High-Frequency Compensation

High-frequency compensation is even more important than low-frequency compensation because of the higher band of frequencies to be passed. In video amplifiers, frequencies as high as 4 megacycles have to be passed with a fairly flat response from the minimum frequencies on upward. Therefore, this type of compensation is considered as the more important one.

There is really only one part of the RC coupling circuit that reduces the response at high frequencies, as compared with the three different networks affecting the low frequencies. Referring to Fig. 7-3, the total shunt capacities, represented by the capacitance C_t, existing in the RC circuit are the chief reasons for the high-frequency response falling off.

Since these shunting capacities are difficult to reduce due to their inherent qualities, compensation must be made to offset their effects. Since we are dealing with very high frequencies, the reactance of these shunting capacities are very low and they offer a low impedance path to these frequencies. Considering a single stage of coupling, these capacities include the output and input capacity of the first and second tubes respectively, the interelectrode capacities of the tubes themselves, and any stray wiring capacities, such as those between the coupling capacitor C_b and ground. The equation for the gain at high frequencies of RC coupling, assuming $R_p>>R_b$ and $R_c>>R_b$, becomes:

Eq. 7-6

$$A_{high} = \frac{g_m R_b}{\sqrt{1+\left(\frac{R_b}{X_{ct}}\right)^2}} = \frac{g_m R_b}{\sqrt{1+(2\pi f C_t R_b)^2}}$$

Let us consider equation 7-6 for a moment. What this relation states (every mathematical expression is a form of statement) is that the gain of an RC coupled amplifier stage at high frequencies, A_{high}, is dependent upon four distinct factors, these are the transconductance or g_m of the tube used, the plate load, or R_b, on the same tube; the frequency of operation f, and the total shunting capacitances C_t. The other quantities used in the expression are no more than actual numbers or numerical constants. The expression further states that the true gain of the RC coupled stage involves the four above mentioned quantities in such a way that the gain is equal to the product of the transconductance of, and the plate load on, the amplifier tube, all divided by the relation inside the square-root sign. This square-root relation is equal to one (1) plus the square of the product of a numerical constant 2π which equals 2×3.14 or 6.28. multiplied by the frequency of operation f, the shunting capacitances C_t, and the plate load R_b.

Consequently from the above expression it is found that if the g_m of the tube in question is high, so will the gain, A_{high}, be high; but if the shunt capacitances, C_t, of the tube and the frequency of operation, f, are high. then the gain A_{high} will decrease. It can be seen how much information this mathematical expression gives. Even an actual computation is not difficult, when all the values of g_m, R_b, f, and C_t are known; it can be easily and simply executed in a few steps. Let us assign values to the symbols and illustrate a typical calculation, which consists of a series of multiplications and a division.

Assuming R_b equals 50,000 ohms, g_m equals 1000 micromhos, f equals 1500 kc, and C_t equals 30 $\mu\mu f$, then:

$$A_{high} = \frac{.001 \times 50{,}000}{\sqrt{1+(2 \times 3.14 \times 1500 \times 10^3 \times 30 \times 10^{-12} \times 50{,}000)^2}}$$

Starting with the denominator (the expression under the square-root sign), we find that $(2 \times 3.14 \times 1500 \times 10^3 \times 30 \times 10^{-12} \times 50{,}000)^2$ is nothing more than a series of multiplications and is equal to 14.1. Squaring this product gives 199. Adding this number to one (1), gives the total number under the square-root sign which is $1 + 199$ or 200. Evaluating the square root of 200, it is found that 14.14 is close enough for practical purposes. Up to this point, only the denominator which is equal to 14.14 has been calculated. The numerator of the expression is equal to $0.001 \times 50{,}000$ or 50. Only one more step is required to obtain the final value of A_{high}, and that is to divide the numerator by the denominator. Therefore:

$$A_{high} = \frac{50}{14.14} = 3.53$$

From the above mathematical computation the gain at the middle frequencies which is equal to $g_m R_b$ is 50, as compared to the gain at the high-frequency end of the band, which is 3.53. This drop in gain, as mentioned before, is due to C_t.

It is seen that with a tube of good mutual conductance and a circuit with the total shunt capacitances C_t small, then the gain at high frequencies will be high. The g_m of the tube is controlled by using special high g_m pentodes designed especially for video-band amplifiers. The undesired effect of C_t is reduced by inserting inductances in the circuit at special places. These inductances effectively increase the load impedance on the amplifier, thereby increasing the gain lost by the shunting capacities. This is analogous to low-frequency compensation where the effective load impedance is also increased. The simplest type of compensation at the high frequencies is to insert an inductance L_p (known as the "peaking coil") in series with the plate-load resistor R_b. L_p is known as a peaking coil because it effectively "peaks" the high-frequency end of the response curve. This type of stage is sometimes referred to as a "shunt-peaked" stage because the coil is placed in shunt with the tube (effectively from plate to ground).

Fig. 7-6 shows a typical video amplifier compensated at the high-frequency end only and its equiva-

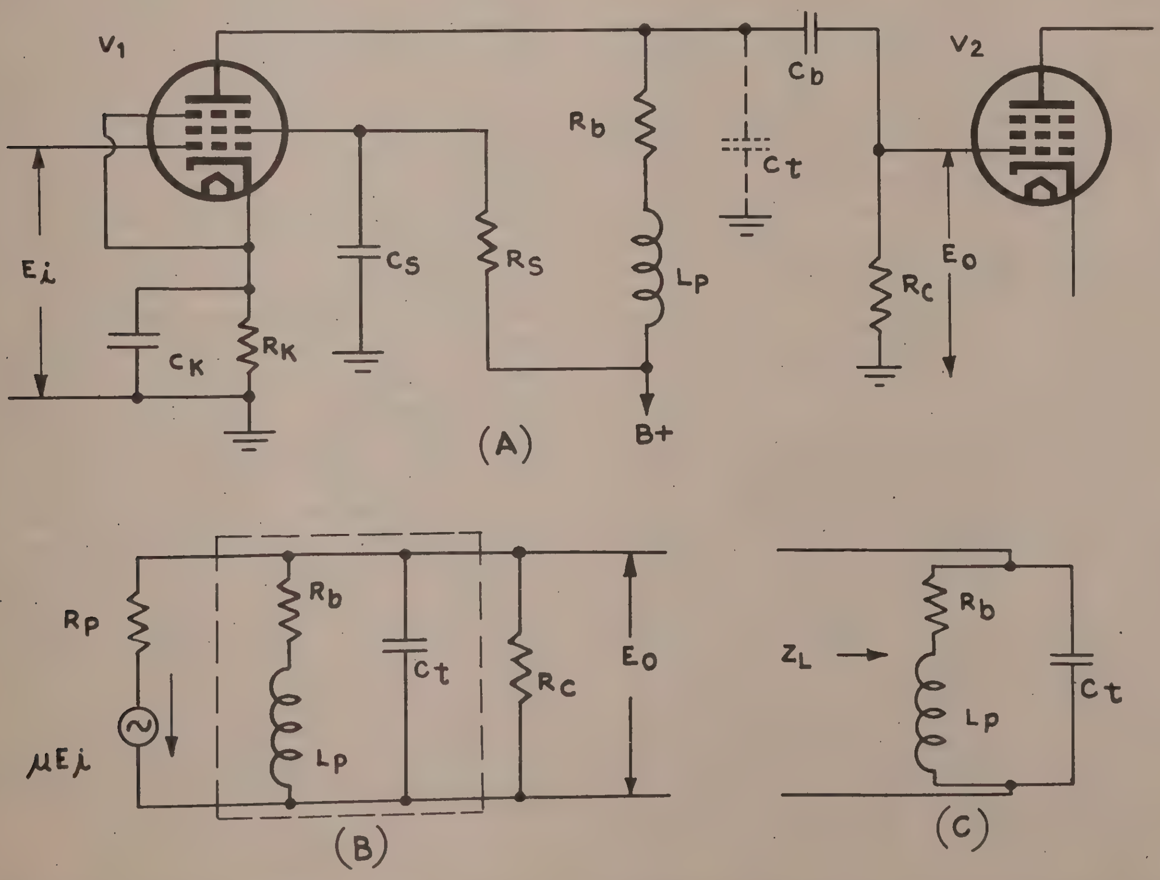


FIG. 7-6.—A typical video amplifier compensated only at the high-frequency end is shown in (A) and its equivalent circuit in (B) and (C).

lent circuit at these frequencies. In Fig. 7-6(B) the coupling capacitor C_b is omitted because at these high frequencies its reactance is so low that it becomes effectively a short circuit. The total load impedance on the first tube V_1 is the combination of R_b in series with L_p, both in parallel with the shunting capacities C_t as shown in the dotted portion of Fig. 7-6(B). An analysis of this circuit is necessary in order to show how the effective load impedance is increased in order to increase the gain. (The grid-leak resistor R_c is neglected as part of the load impedance because it is usually very high and it has little effect on the parallel branch.)

The total load impedance represented by Z_l is given as follows:

$$Z_l = \frac{X_{ct}\sqrt{R_b^2 + X_{lp}^2}}{\sqrt{R_b^2 + (X_{lp} - X_{ct})^2}} \qquad \text{Eq. 7-7}$$

It is known that a parallel resonant circuit has a very high impedance at resonance and, therefore, a large voltage will be developed across it. Consequently, since the effective load impedance has to be increased to offset the drop in response, the parallel circuit of R_b, L_p and C_t is made to resonate at the highest frequency that we desire to amplify. The proper values needed for this resonance are chosen from R_b and L_p, because C_t is fixed. Assuming that the resonant frequency has already been established in the circuit, the capacitive reactance is equal to the inductive reactance ($X_{Lp} = X_{Ct}$), provided that the resistance R_b has a minor effect. The load impedance (Eq. 7-7) then becomes:

$$Z_l = X_{ct}\sqrt{1 + (X_{lp}/R_b)^2} \qquad \text{Eq. 7-8 (A)}$$

or $$Z_l = X_{ct}\sqrt{1 + (X_{ct}/R_b)^2} \qquad \text{Eq. 7-8 (B)}$$

or $$Z_l = X_{lp}\sqrt{1 + (X_{ct}/R_b)^2} \qquad \text{Eq. 7-8 (C)}$$

or $$Z_l \doteq X_{lp}\sqrt{1 + (X_{lp}/R_b)^2} \qquad \text{Eq. 7-8 (D)}$$

Eq. 7-8 (A), (B), (C) and (D) are all equal because they are evaluated at the resonant frequency where $X_{Lp} = X_{Ct}$ under the assumption above.

From our previous knowledge of resonance curves, it is known that the critical frequency or the minimum frequency (for either the high or low end) acceptable for design purposes is that point on the curve where there is a 3 db drop in gain. At the high end, it is this frequency that we start with in correcting the gain because all the frequencies between the intermediate ones and this high one fall into the acceptable area. For good response in *RC* coupled amplifiers the load resistance is made equal to the reactance of its shunting capacities, which is what we have at the 3 db point. Therefore we get:

$$R_b = X_{ct} = \frac{1}{2\,\pi\, f_c\, C_t} \qquad \text{Eq. 7-9}$$

where f_c = the frequency where correction starts.

Under this criterion the radical of equation 7-8 becomes equal to the square root of two and the minimum load impedance that will be presented to the amplifier at the resonant frequency of L_p and C_t is:

$$Z_l = X_{ct}\sqrt{2} = 1.41\, X_{ct} \qquad \text{Eq. 7-10 (A)}$$

or $$Z_l = X_{lp}\sqrt{2} = 1.41\, X_{lp} \qquad \text{Eq. 7-10 (B)}$$

We must decide what is the maximum frequency that we desire to compensate for. Assuming this frequency to be the 4 megacycles mentioned, then this 4 megacycle frequency is known as the correction frequency for which the wide band video amplifier must be designed. When the parallel impedance becomes effectively a pure resistive load to the tube (at resonance), then the total reactance of the circuit is zero and the reactance of one branch is equal to that of the other. At this point the inductance L_p becomes:

$$L_p = C_t R^2_b \qquad \text{Eq. 7-11}$$

Eq. 7-11 states that if we know the load resistance R_b used and have figured the shunting capacities, we can calculate the value of shunt peaking inductance (L_p) needed. It has been found from practice that when a response curve is plotted for all values of frequency versus gain, using the compensated circuit of Fig. 7-6(C) and Eq. 7-11, the high-frequency end of the curve is improved considerably over that of an ordinary *RC* coupled circuit. The amplitude or the gain of the response curve receives the improvement, but the over-all curve is not as flat as we would desire because the value of inductance used (as calculated by Eq. 7-11) seems to be too large and has caused too high a rise in gain at the critical frequency. The rise or peak of the curve due to coil L_p is the reason that L_p is termed a "peaking" coil. From experimentation it has been found that for over-all flat response, down to the frequency where the load resistance R_b is equal to the capacitive reactance of C_t, the peaking coil should have a value equal to *half* of that represented by Eq. 7-11 or mathematically:

$$L_p = 1/2\, C_t (R_b)^2 \qquad \text{Eq. 7-12 (A)}$$

and since we know that at the frequency of correction

$$R_b = \frac{1}{2\pi f_c C_t} \quad \text{where } C_t = \frac{1}{2\pi f_c R_b}$$

and substituting this in Eq. 7-12 (A) we get

$$L_p = 1/2 \times \frac{1}{2\pi f_c R_b} \times (R_b)^2 = \frac{0.5 R_b}{2\pi f_c}$$

and since

$$\omega_c = 2\pi f_c$$

then L_p can be written as

$$L_p = \frac{0.5\ R_b}{\omega_c} \qquad \text{Eq. 7-12 (B)}$$

Eq. 7-12 (A) and (B) above are design formulas that have actually been derived from practice. The response curve will be very flat down to the frequency of correction if the above formulas are used. The actual resonant frequency of the parallel circuit (in which L_p is derived by formula 7-12) is approximately $\sqrt{2}$ or 1.41 times the frequency of correction.

In summarizing the above equations we can say that the shunt peaking inductance L_p can be determined in one of two ways.

First, all that would have to be known to find L_p would be load resistance R_b and the frequency of correction and not the shunting capacities. This is evident from Eq. 7-12 (B).

Secondly, we do not have to know the frequency of correction in order to find the shunt peaking inductance L_p but just the load resistance R_b and shunting capacities C_t. This is evident from Eq. 7-12(A).

As an example, suppose that the shunting capacities of a stage such as is shown in Fig. 7-6(A) are found to be equal to 24 $\mu\mu f$ measured at a frequency of 3 megacycles. We must now determine the value of the shunt peaking coil L_p and the value of the plate-load resistor R_b needed. We know that for flat response at the high frequencies, the plate-load resistor R_b must equal the reactance of the shunting capacities at the maximum frequency to be amplified (which then becomes the correction frequency). Therefore, with $f_c = 3$ megacycles:

$$R_b = \frac{1}{2\pi f_c C_t} = \frac{1}{2\pi\ (3 \times 10^6)\ (24 \times 10^{-12})}$$
$$= 2210 \text{ ohms}$$

and for the peaking coil L_p, from Eq. 7-12 (A) and (B):

$$L_p = \tfrac{1}{2} C_t R^2{}_b$$
$$= \tfrac{1}{2} \times 24 \times 10^{-12} \times 2210$$
$$= 58.6 \text{ microhenrys}$$

or

$$L_p = \frac{0.5 R_b}{2\pi f_c} = \frac{2210}{4\pi\ (3 \times 10^6)} = 58.6 \text{ microhenrys}$$

Therefore we find that the *approximate* value needed for the plate-load resistance R_b is about 2200 ohms and that for the shunt peaking inductance L_p is about 60 microhenrys.

As an illustration of how the peaking inductance L_p affects the high frequencies let us examine some typical response curves. Fig. 7-7 shows a number of different curves, where each one utilizes a different value of L_p. Note that the curve where the response is the flattest is when $L_p = 1/2\ C_t R^2{}_b$.

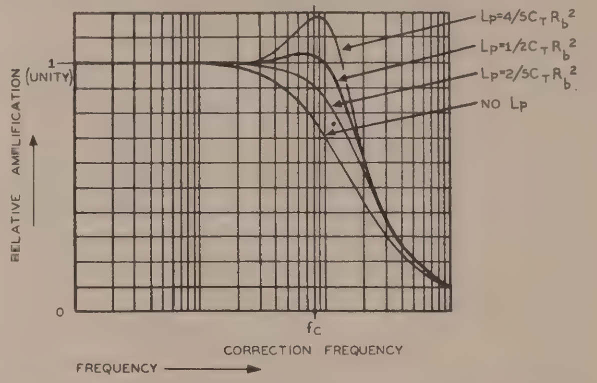


FIG. 7-7.—The manner in which the peaking inductance affects the high frequencies is shown in these typical response curves, each utilizing a different value or L_p.

Series Peaking Compensation

Another type of high-frequency compensation is known as series peaking. This type of circuit is shown in Fig. 7-8 (A) and (B). Here the peaking coil is placed in series with the plate of the amplifier, instead of in parallel with it. This type of compensation is used when higher gain and more linear phase shift are desired than a shunt compensated circuit provides. (In the previous case the phase shift is also compensated but not in a linear fashion.) Fig. 7-8(A) shows a typical "series peaking" high-frequency compensated video amplifier circuit. In this type circuit the peaking coil L_s (it is still called a peaking coil because it still "peaks" up the high-frequency end of the response curve) is inserted in such a way that it isolates the shunting capacities related to both tubes. C_1 represents those shunting capacities relative to tube V_1 and C_2 represents those shunting capacities relative to tube V_2. The combination of C_1, L_s and L_2 represents a "filter" network which has the characteristics of a low-pass filter (wide-band). That is why series peaking compensation is sometimes known as "filter" coupling. By a "filter" we mean a circuit so designed that it will pass certain frequencies and attenuate others that are undesired.

Variations in the circuit of Fig. 7-8(A) are sometimes made in order to isolate the shunting capacities at some specific ratio. For instance, in Fig. 7-8 (B) the series peaking coil L_s is shifted to the other side of the load resistor because in some circuits the shunted capacities are more evenly distributed under these circumstances. While in Fig. 7-8(C), the coupling capacitor C_b is placed on the left hand side of the peaking coil L_s, so that the shunting capacity between the capacitor C_b and ground will be added to C_1 and not to C_2 as in Fig. 7-8 (A). In all of the above cases, the series peaking coil is inserted in such a way that the shunting capacities are divided on either side of the coil with the ratio between C_1 and C_2 that is most beneficial to the circuit. For most practical purposes and in our discussion C_2 is equal to twice C_1. In practice the load resistor R_b is usually placed on the side of the lower shunting capacity, namely C_1, however, [as seen in Fig. 7-8 (B)] this is not always done since the capacities may be so nearly equal that it makes little difference. Since the total shunting capacities are isolated by the coil L_s, the value of the load resistor R_b in most cases depends upon C_1; whereas the value of the load resistor R_b in shunt peaking depended upon the total capacity C_t. Therefore, the lower the value of C_1, the higher the value R_b can be, and the greater the gain, since the gain depends upon the size of R_b. It then follows, from the above discussion, that both the values of C_1 and C_2 have to be estimated quite closely for the proper design and use of the plate-load resistor R_b and the series peaking coil L_s.

The addition of the series peaking coil helps the high frequency response in the following way [referring to the typical circuit of Fig. 7-8 (A)]: The voltage output from tube V_1 is impressed across the plate-load resistor R_b. Assuming no compensation (L_s removed), then this voltage across R_b would be attenuated at the high frequencies due to the effect of the total shunting capacity. With peaking coil L_s inserted, the total shunting capacities are divided, with the lesser amount going to C_1. The previous loss which was due to the effect of the total shunting capacities is now greatly reduced by the effect of C_1 alone. This output voltage from tube V_1, that is now developed across R_b and C_1, in parallel, is impressed on the voltage dividing network of L_s and C_2. C_b presents a short circuit at high frequencies and the value of the grid-leak resistor R_c is so large compared to the reactances of L_s and C_2 that they (C_b and R_c) can be considered as having negligible effect on the voltage dividing net-

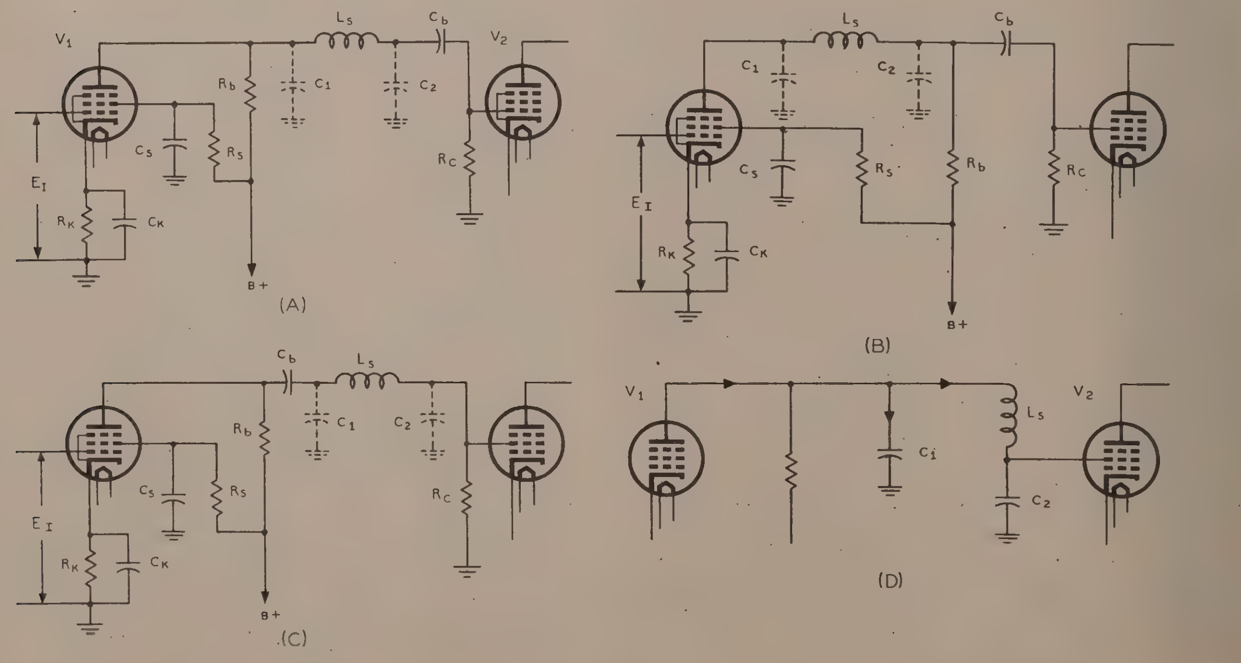


FIG. 7-8.—A typical series peaking high-frequency compensated video amplifier circuit is shown in (A) and a simplification of this circuit in (D). The circuits in (B) and (C) are variations of (A), these being sometimes used to isolate the shunting capacities at some specific ratio.

work L_s and C_2, and are consequently neglected in the circuit analysis at high frequencies. Even though the shunting capacity across tube V_1 is reduced from the total shunting capacity C_t to just the shunting capacity of C_1, there still remains some loss at high frequencies due to this C_1. The problem now is to offset the high-frequency attenuation caused by C_1. This can be done by offering an impedance to the output voltage of tube V_1 at high frequencies that is considerably *lower* than that represented by the R_bC_1 combination, so that most of this signal output voltage will follow the path of low impedance and very little will be shunted to ground through C_1.

The question that now arises is how this effect can be attained. Referring to Fig. 7-8(D), which is a simplified circuit of Fig. 7-8 (A); at some frequency, call it f_c, we notice that L_s and C_2 are in series resonance. Therefore, at this frequency the impedance offered by the series resonant circuit of L_sC_2 to the output of tube V_1 will be a minimum, and *most* of the output signal voltage from tube V_1 will follow the path of minimum impedance offered to the tube, namely L_sC_2, and very little will be shunted to ground by the higher impedance of C_1. The signal current flowing from tube V_1 will then flow through the voltage dividing network of L_s and C_2, and the maximum amount of voltage (at high frequencies) will be impressed on the grid of tube V_2 by the voltage drop across capacitor C_2. It is, therefore, seen how capacitor C_2 offsets the drop in gain at high frequencies by the characteristics of its resonant effects, at these frequencies, with the series peaking inductance L_s. In actual practice, the ratio between C_2 and C_1 is set approximately equal to two, as stated before. The values of C_1 can be measured pretty accurately and so can the total capacitance C_t ($C_t = C_1 + C_2$), and from knowing C_1 and C_t, C_2 can be found. Circuit components may have to be shifted around in order to find the approximate ratio.

In working with design formulas for the series peaking inductance and also for the load resistance, the operation of the series peaking circuit of Fig. 7-8(A) may be better understood by discussing typical response characteristics at the high end of a curve—such as shown in Fig. 7-9.

In this figure the gain is only relative and its maximum value is taken as unity at the intermediate frequencies.

Curve 1 represents the actual curve of the amplifier without any compensation. It is readily seen how this curve falls off at the high frequencies. The horizontal dashed curve 3 is the 3 db curve where the gain drops to 70.7% of its maximum value. (Keep in mind that this 3 db value is the point of minimum gain allowable for most proper designs.) Assume that we want to have a flat frequency response characteristic to a correction frequency f_c of 3 mc. Then, from curve 1 we note that at this frequency the gain is at point (*a*) and therefore undesirable. We want to bring point (*a*) up to somewhere in the shaded area. Ideally we want to bring point (*a*) up in such a manner that curve 1 is raised at the high frequencies so it has a relative voltage gain of unity. In other words, even though upon correction, point (*a*) may advance anywhere into the shaded area (the area of allowable gain) between points (*b*) and (*c*), it is preferable to approach point (*c*). By actual experimental determination, it has been found (when $C_2 = 2C_1$) that this correction is ideally approached when the elements L_s and C_2 are made to resonate at the correction frequency f_c. Under such conditions curve 1 is shifted to curve 2. From this curve it is evident that at the correction frequency f_c, there is a peak rising *slightly* (which is due to series resonance between L_s and C_2) above the level of unity gain; but in general the shape of the curve approaches the ideal case very closely. We therefore have two ways to explain series peaking compensation—one by the explanation of the actual response curve and the other by the theoretical explanation of how the low impedance path due to L_sC_2 helps the gain at high frequencies.

The formulas for the design of the series peaking coil L_s and the plate-load resistance R_b depend primarily on the value of the frequency desired for correction. Calling this frequency, as we have before, the correction frequency f_c, we know now that L_sC_2 should be made to resonate at this frequency, whereupon we have:

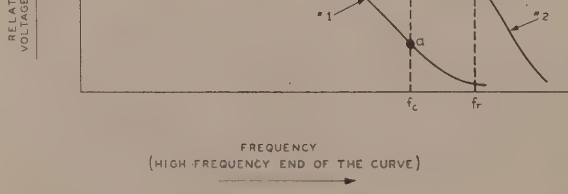


FIG. 7-9.—Typical response characteristic curves, wherein the voltage gain is relative, its maximum value being taken as unity at the intermediate frequencies.

$$f_c = \frac{1}{2\pi\sqrt{L_s C_2}} \qquad \text{Eq. 7-13 (A)}$$

where the shunting capacity C_2 is known, by measurement as discussed beforehand, or by close approximations. Squaring both sides of the equation above and solving for L_s we get:

$$L_s = \frac{1}{(2\pi f_c)^2 C_2} \qquad \text{Eq. 7-13 (B)}$$

and also knowing that $C_2 = 2C_1$ and substituting this in Eq. 7-15(B) we get:

$$L_s = \frac{1}{(2\pi f_c)^2 2C_1} \qquad \text{Eq. 7-13 (C)}$$

From Eq. 7-15(B) and (C) we can readily solve for the series peaking inductance in terms of either shunting capacity C_1 or C_2. We can rearrange Eq. 7-13(C) in the following manner:

$$(2\pi f_c)\,L_s = \frac{1}{2}\cdot\frac{1}{(2\pi f_c)\,C_1} \qquad \text{Eq. 7-13 (D)}$$

which states that the inductive reactance of L_s at the frequency of correction is equal to 1/2 the capacitive reactance of C_1 at the same frequency.

If we rearrange Eq. 7-13(C) in another way by solving the equation for the value, $\sqrt{2}\,f_c$, we find the following:

Taking the square root of both sides of Eq. 7-13 (C) we have:

$$\sqrt{L_s} = \frac{1}{2\pi f_c\sqrt{2}\sqrt{C_1}}$$

and solving for $\sqrt{2}\,f_c$ we get:

$$\sqrt{2}\,f_c = \frac{1}{2\pi\sqrt{L_s C_1}} \qquad \text{Eq. 7-13 (E)}$$

Referring to equation 7-13 (E), it will be noticed that the right-hand side of the expression is itself representative of a resonant frequency. In other words, a frequency exists at which the series peaking coil L_s is in resonance with the lower valued shunting capacity C_1. The resonant frequency of L_sC_1 is then equal to $\sqrt{2}$ times the frequency of correction. That is to say, if we let f_r equal the resonant frequency of L_s and C_1, then

$$f_r = \sqrt{2}\,f_c \qquad \text{Eq. 7-13(F)}$$

Eq. 7-13(F) means that there exists for L_s and C_1 a resonant frequency (f_r) that is equal to $\sqrt{2}$ times the frequency of correction.

Knowing that the total shunting capacity C_t is equal to the sum of C_1 and C_2 and that

$$C_1 = \frac{C_2}{2}$$

then substituting for C_1, C_t becomes:

$$C_t = \frac{C_2}{2} + C_2 = \frac{3}{2}C_2$$

and solving this equation for C_2 we get:

$$C_2 = 2/3\;C_t$$

and substituting this equation in Eq. 7-15(B) we get:

$$L_s = \frac{3}{(2\pi f_c)^2\,2\,C_t} \qquad \text{Eq. 7-13 (G)}$$

rearranging we get:

$$2\pi f_c L_s = \frac{3}{2}\cdot\frac{1}{2\pi f_c C_t}$$

or

$$X_{ls} = \frac{3}{2}X_{ct} = 1.5\,X_{ct} \qquad \text{Eq. 7-13 (H)}$$

which states that the inductive reactance of the series peaking coil at the frequency of correction is equal to 3/2 the reactance of the total shunting capacity C_t at the frequency of correction.

In determining the value for the inductance of the peaking coil we can utilize any one of the above eight equations [Eq. 7-15(A to H)] according to the parameters that are known. These equations make use of the fact that in order to solve for L_s all that has to be known are any two of the following elements that fit into *the particular formula* used:

C_t = total shunting capacity in the circuit $(C_1 + C_2)$

C_1 = shunting capacities around the output of the first tube

C_2 = shunting capacities around the input of the second tube

f_c = frequency of correction (or that frequency where L_s and C_2 will be in series resonance)

f_r = that frequency which is $\sqrt{2}$ times the frequency of correction (or that frequency where L_s and C_1 can be considered to be in resonance)

Now that we have decided on numerous ways for determining L_s, we are ready for the proper determination of the plate-loading resistor R_b in these series peaking circuits.

From all other previous topics we have determined that the plate-load resistance R_b should be equal to its shunting reactance for proper design purposes. In shunt peaking, this reactance was the total shunting reactance X_{ct} in the circuit.

Therefore, for design purposes R_b was set equal to $1/X_{ct}$ at the frequency of correction, which was the minimum acceptable 3-db frequency. One should understand that these assumptions are not just guesses but are based upon actual experimental data taken on such amplifier circuits. When the response curve of Fig. 7-9 was plotted, it was found by experimentation that the plate-load resistor R_b was approximately equal to *one-half* the reactance of the shunting capacitor C_1 at the frequency of correction. Accordingly, the basic design formula for the load resistor R_b for series peaking is:

$$R_b = \frac{1}{2}X_{c1} = \frac{1}{2\,(2\,\pi\,f_c\,C_1)} \qquad \text{Eq. 7-14 (A)}$$

Since it is easier to measure the total shunting capacities C_t than C_1 alone, we shall rearrange Eq. 7-14(A) in terms of C_t instead of C_1 by substituting

$$C_1 = \frac{C_t}{3} \text{ (since } C_2 = 2C_1\text{)}$$

Hence

$$R_b = \frac{3}{2}\cdot\frac{1}{2\,\pi\,f_c\,C_t} = \frac{3}{2}\,X_{ct} = 1.5\,X_{ct} \qquad \text{Eq. 7-14 (B)}$$

Eq. 7-14 (B) states that for proper series peaking compensation, the plate-load resistor R_b is made equal to one and one-half times the total shunting reactance X_{ct} at the frequency of correction (when $C_2 = 2C_1$).

We also can combine the relation between the equations 7-13(D) and 7-14(A) in such a manner that L_s and R_b are in the same equations. Substituting Eq. 7-14 (A) into Eq. 7-13(D) we get:

$$X_{ls} = R_b$$

and substituting $2\pi f_c L_s$ for X_{LS} we get:

$$2\,\pi\,f_c\,L_s = R_b$$

where

$$L_s = \frac{R_b}{2\,\pi\,f_c} \qquad \text{Eq. 7-15 (A)}$$

Solving Eq. 7-14(B) for $2\pi f_c$ we get:

$$2\,\pi\,f_c = \frac{3}{2\,C_t\,R_b}$$

and substituting this equation in Eq. 7-15(A) we get:

$$L_s = \frac{R_b}{3/(2\,C_t\,R_b)} = \frac{2}{3}\,C_t\,(R_b)^2 = .67\,C_t\,(R_b)^2 \qquad \text{Eq. 7-15(B)}$$

From the above equations it is seen that we can solve for the plate-load resistance R_b by knowing the frequency of correction f_c and any one of the following quantities—C_1, C_t, or L_s.

Some fundamental properties of the series peaking circuit are that it has 50 percent more gain than a shunt peaking circuit (compare Eq. 7-14 with Eq. 7-9) where both circuits have the same values of R_b, f_c and C_t; and also that the phase shift response in series peaking is more linear than in shunt peaking. Thus it is seen that series peaking betters the gain and phase characteristics of the amplifier at the high-frequency end of the curve as compared to shunt peaking.

To further better the amplification and phase shift characteristics, two such series peaking coils are put into a circuit as is shown in Fig. 7-10. This type of circuit improves the response characteristics that one such series peaking network gives and also provides some other terminals from which extra signal energy may be derived for uses such as synchronization.

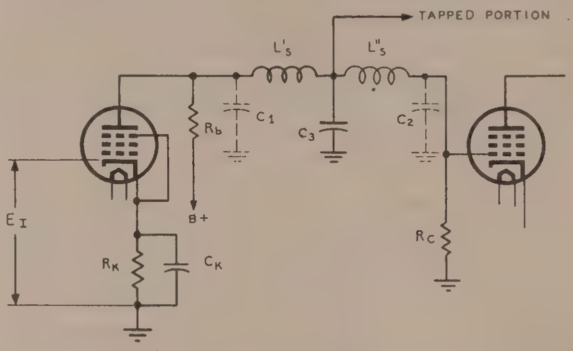


FIG. 7-10.—In order to improve amplification and phase-shift characteristics, two series peaking coils are employed, thus bettering the response characteristics that one such coil provides.

So far we have studied two separate high-frequency compensating networks, namely shunt peaking and series peaking. The latter type of peaking was found to give a better response and phase shift than the former, but a combination of these types gives a better response than either one alone. Fig. 7-11 illustrates a typical circuit combining the effect of series and shunt peaking with the shunting capacities C_1 and C_2 shown. With this type of compensation there is a sharper cutoff at the correction frequency than in the series case alone. Much greater gain is obtained than when shunt or series peaking alone is used. It has been found that combination peaking gives as much as 80 percent more gain than a shunt peaking system alone. This means that the total effective load impedance is also 80 percent greater than that for shunt peaking.

So far we have discussed the separate types of frequency compensation (high and low) which may be applied separately to any system desired. But in video amplifiers, as previously pointed out, the desire is to have as flat a response as possible from the lowest audio frequency encountered to the highest video frequency needed. It is evident that to obtain these qualities in a single circuit, the low-frequency and high-frequency compensated networks have to be combined. Fig. 7-12 illustrates a number of different video amplifiers in which both low- and high-frequency compensation are employed. An interesting feature of these circuits is the additional capacitors used.

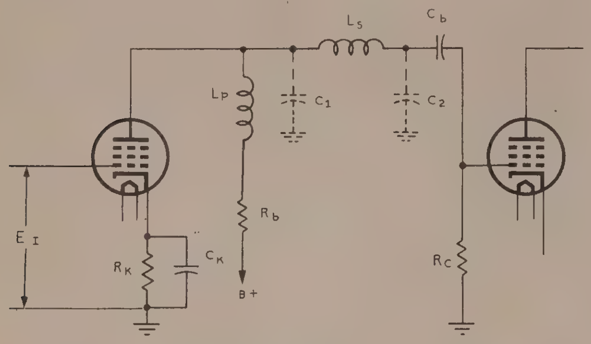


FIG. 7-11.—A typical circuit combining the effect of series and shunt peaking with the shunting capacities C_1 and C_2 indicated across the coil L_s.

In all the circuits it will be noted that the capacitors C'_k, C'_s, and C''_b are shunted across the capacitors C_k, C_s, and C'_b that were originally used in the low-frequency compensated case (see Fig. 7-4). One of the fundamental actions of capacitors C_k, C_s, and C'_b is to bypass low or audio-frequency components to ground and, therefore, prevent these frequencies from getting into certain parts of the circuit where they are not desired. For instance, feedback is prevented by such bypassing as was previously discussed. The values of these capacitors C_k, C_s, and C'_b are hence made large so that their reactance to low frequencies will be small. That is fine for low frequencies; but when the low-frequency compensated network is combined with the high-frequency compensated network, a separate path to bypass the high frequencies must be provided to prevent feedback. Two separate bypassing paths are thus needed, one for low frequencies and the other for the high. We already have the low-frequency bypass capacitors in the circuit, and since bypassing of the high frequencies is necessary, then all that has to be done is to shunt capacitors C_k, C_s, and C'_b with the other capacitors C'_k, C'_s, and C''_b, respectively. Since these latter three capacitors are needed to bypass the high frequencies they are much smaller than the former ones. Some typical values employed in practice are as follows:

$C_k = 100\ \mu f$ $\qquad C_s = 16\ \mu f$ $\qquad C'_b = 30\ \mu f$

$C'_k = 0.1\ \mu f$ $\qquad C'_s = 0.01\ \mu f$ $\qquad C''_b = 0.1\ \mu f$

In visualizing these capacitors one must remember that their size depends not only on the capacity value but also on the voltage rating. For example,

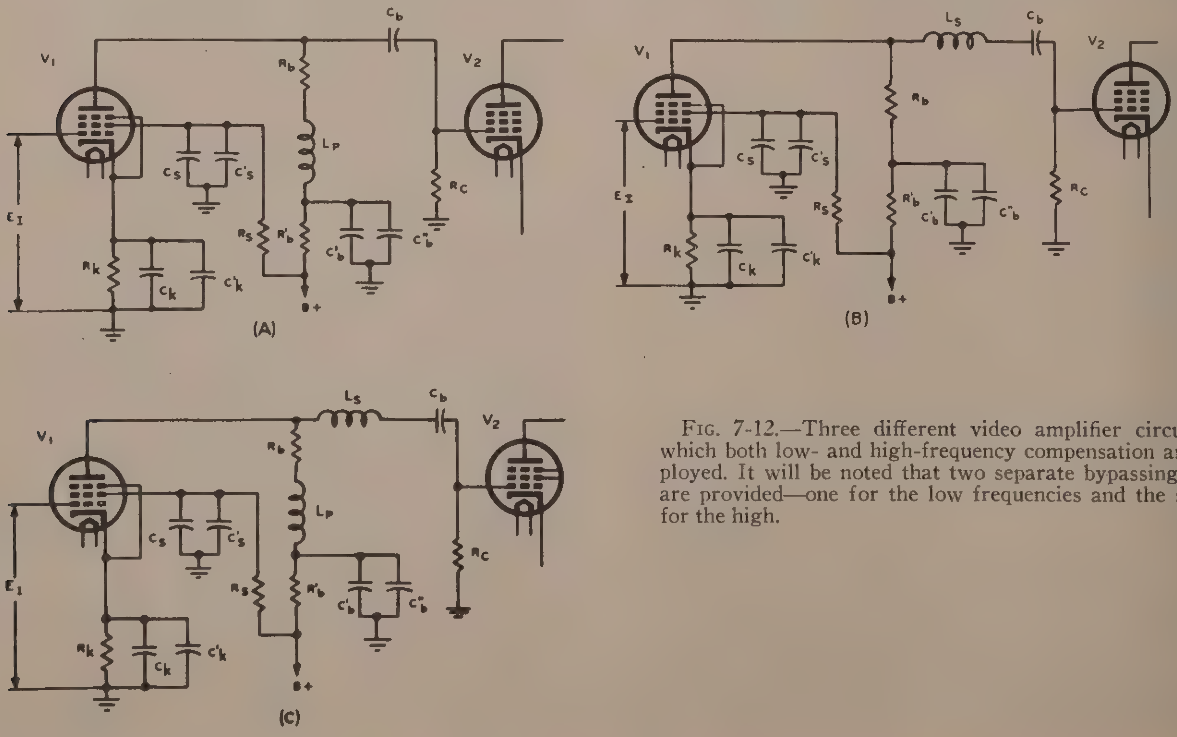


FIG. 7-12.—Three different video amplifier circuits in which both low- and high-frequency compensation are employed. It will be noted that two separate bypassing paths are provided—one for the low frequencies and the second for the high.

the 100 µf cathode bypass capacitor need only have a small value of breakdown voltage depending on the voltage drop across the cathode resistor R_k.

Assuming the bias on the tube to be —10 volts, then the drop across the bias resistor R_k is 10 volts and for a factor of safety the cathode bypass capacitor C_k (and likewise C'_k) usually has a voltage rating of 25 or 50 volts. Such a capacitor of 100 µf has to be of the electrolytic type and it is not very large, since the breakdown voltage is to be 25 or 50 volts.

RCA Model 630TS

The first video amplifier of the RCA model 630TS receiver is shown in Fig. 7-13. The coupling network between the first video amplifier and the second video amplifier employs both low- and high-frequency compensation to produce a flat response and linear phase shift over the entire range of video frequencies, approximately 30 cycles to 4 megacycles. The low-frequency compensation is accomplished by resistor $R141$ and capacitor $C223B$ in the plate circuit. This network corresponds to R'_b and C'_b in the previous discussion of low-frequency compensation. The plate-load resistor $R140$ corresponds to R_b, the coupling capacitor corresponds to C_b, and the grid-load resistor R_c would be the equivalent resistance of $R142$ in parallel with $R143$, which is 487,000 ohms. If the compensating circuit were not used, the gain at low frequencies would fall off due to the effect of the $C140$—$R142$, 143 network. The reactance of capacitor $C140$ goes up as the frequency goes down, causing less of the output voltage of the first stage to appear at the grid of the second stage. To correct for this, $R141$ and $C223B$ were chosen so that the impedance of this combination would increase with lower frequencies in the proper ratio to compensate for the falling off due to the coupling circuit to the next stage. This compensation occurs because a higher load impedance in the plate circuit of the 6AU6 results in a greater output signal.

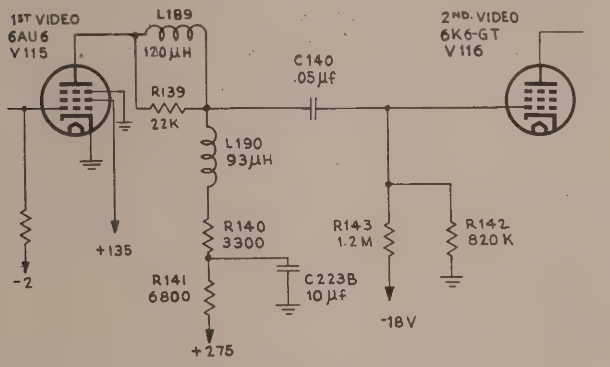


FIG. 7-13.—The first video amplifier circuit of the RCA model 630TS receiver.

Courtesy RCA

The values of resistance and capacitance are also chosen to make the phase shift linear with respect to frequency. The value of $R141$ is limited by the drop in voltage across this resistor, which decreases the plate voltage on the 6AU6. It is interesting to note the time constant of the $R141$—$C223B$ combination. This time constant is equal to 6800×10^{-5} or 0.068 second. Referring to Fig. 7-5 which shows the amplitude response for this type of low-frequency compensation and interpolating between the curves shown, it can be seen that under ideal conditions the response will be flat to about 10 cycles, which is well below the lowest frequency (30 cycles) that has to be passed.

The high-frequency combination employs both series and shunt peaking. This is done so that the response at the high frequencies will be better than if only shunt or series peaking were used. The shunt peaking used in this circuit is similar to that shown in Fig. 7-6(A). The peaking coil $L190$ is equivalent to the peaking coil L_p in the previous discussion. This coil resonates with the shunt capacitance across the input circuit of the 6K6. This includes the input capacitance of the 6K6, about 5.5 µµf, plus all the stray capacitance due to the wiring and the parts used in this circuit. It is difficult to estimate this capacitance as it varies quite a bit, depending on the method of wiring used, the physical size of each component, its proximity to ground, etc. In the actual design of a circuit like this, the shunt capacitances are actually measured with an impedance bridge or similar instrument and then the correct values of the peaking coils and the plate-load resistance are determined from the design equations that have been discussed.

The series peaking coil $L189$ is equivalent to the peaking coil L_s in our previous discussion of series peaking. The output capacitance of the 6AU6, which is about 5 µµf, plus the stray capacitance in this circuit correspond to the shunting capacitance C_1, and the input capacitance of the 6K6 plus strays correspond to the shunting capacitance C_2 in Fig. 7-8 (B). The coil resonates with these shunting capacitances to produce a peak at the high end of the frequency range. This peak, together with the peak caused by the shunt-peaking coil, is sufficient to flatten the response up to the highest video frequency, about 4 megacycles. The 22,000-ohm resistor, $R139$, is placed across the peaking coil $L189$ in order to reduce the effective Q of the coil. It was probably found that the resonant peak produced by this coil in conjunction with the shunt capacitance was too high. To get an over-all frequency response that is fairly flat over the entire range, it was necessary to reduce this peak. This can be done by damping, or lowering the Q of the resonant circuit, and is accomplished in this particular circuit by placing the resistor $R139$ in shunt with the series-peaking coil, $L189$. In a circuit that uses both series and shunt peaking, the shunt-peaking and the series-peaking coils may both be used to peak at the same frequency if the loss at some particular high frequency is quite high. Or, the resonant frequency between both circuits may differ somewhat, so that their resonant peaks may be slightly apart. This results in an increase in voltage gain at a wider range of high frequencies than if both coils are peaked at the same frequency.

AVERAGE BRIGHTNESS AND D-C RESTORER CIRCUITS

In our previous discussion of the video signal, we looked upon this signal as containing a series of voltage values each of which corresponded to a particular value of light intensity in the televised image. We considered that in the same way that light values can be reckoned from black as a reference level, so a particular voltage value can be assigned to black and then each light value represented electrically by assigning a higher (or lower) voltage to the signal, depending upon the brightness of the light value at the scanned area. This method of looking at a video signal is shown in Fig. 7-14 (A), in which it is clear that all light values are with reference to the black level—which is taken as the zero-voltage axis.

Insofar as video amplifier operation is concerned, the important thing about Fig. 7-14(A) is that it shows that a video signal is inherently a pulsating d-c voltage. In other words (as in the case of a rectified a-c voltage which is also a pulsating voltage) all the light values are represented by electrical values on one side of the zero-voltage axis. The video signal must therefore contain both a d-c and an a-c component in the same way that a rectified a-c voltage representing the output of a power supply contains a d-c component—the d-c voltage of the power supply—and an a-c component—the hum ripple of the power supply.

Physically what does it mean to say that a video signal contains a d-c as well as an a-c component? Actually it is only another way of looking at the signal, as Fig. 7-14(A) clearly shows. In this figure we describe the signal by saying that the light values are represented by fluctuations in both directions from an average level, which we can call the "average brightness," or the "picture background." Thus instead of describing the light value at any point in the scene by stating how much brighter it is than black, as in (A), in (B) we accomplish exactly the same thing by stating how much brighter or blacker is the particular light value than the *average brightness*. In (A) we use the black level as the reference level whereas in (B) we use the average brightness as the reference level.

Electrically speaking, the average brightness level represents the average value of the video signal, or in other words the d-c component of the signal. On the other hand, the fluctuations in the signal on either side of the average brightness level, which is electrically the a-c axis of the signal, represent the a-c component of the signal. An important characteristic of the average brightness or a-c axis is that the area between the positive part of the cycle and the a-c axis is equal to the area included between the negative part of the cycle and the same a-c axis. This is shown in Fig. 7-14(B), where it can be seen that area 1 is equal to area 2 + area 3.

From the preceding it will be clear that proper operation of the picture tube requires that both the d-c and a-c components of the video signal be passed by the video amplifier. For if only the a-c component reaches the grid of the picture tube, then the picture tube will have information only on the fluctuations in light values *with reference to the average brightness level* but it will have no information whatsoever on the value of this average brightness level. Thus the pic-

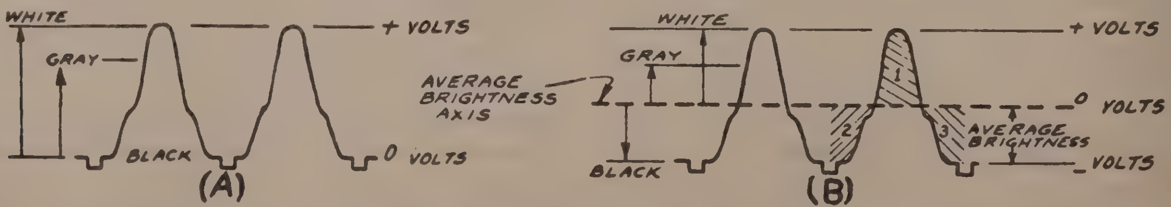


FIG. 7-14.—The light values in a video signal can be reckoned in two ways as here illustrated: in (A) the various light values are referred to the black level; in (B) the same light values are referred to the average brightness of the signal.

ture cannot be reproduced accurately since the same variations (represented by the a-c component) might be superimposed on either a dark or a light background of any shade. The average brightness, or the d-c component of the video signal, must be present before the picture can be reproduced.

D-C Restorer Circuits

Unfortunately, it is not possible to transmit the d-c component of any signal without using a direct-coupled amplifier. However, the use of direct-coupled amplifiers is not practical in television receivers because of their comparative instability and high cost. For this reason, television engineers have developed circuits which make possible the use of conventional a-c amplifiers with capacitor coupling and at the same time provide for the restoration or *recreation* of the d-c component after the a-c component has been amplified by itself.

It will be helpful at this time to consider several video signals which have different values of average brightness and to describe the action which takes place when these signals pass through an amplifier which employs a blocking capacitor as a coupling element between stages. It is because of the presence of the blocking capacitor that only the a-c component is passed and the d-c component is lost.

In Fig. 7-15, the picture being scanned is a white triangle on a black background, with the vertex of the triangle near the top. Part (A) shows two lines scanned near the top of the picture; part (B), near the middle of the picture; and part (C), near the bottom. Thus (A), (B), and (C) represent three sections of the picture where the average brightness is low, medium, and high respectively. Fig. 7-15 can be said to represent the signal as it is at the output of the camera tube or as it appears at the video second detector. Of special importance is the fact that all the blacks are lined up so that the d-c component of the signal is represented in all cases. Note that the d-c component, like the average brightness which it represents, is successively larger in (A), (B), and (C).

Now what happens to this signal when it is passed through an a-c amplifier which contains coupling capacitors and, therefore, will not pass the d-c component? This is clearly illustrated in Fig. 7-16 which shows that the blacks are no longer lined up, but instead the separate average-brightness or a-c axes are all lined up. In other words, when the d-c component is lost, only the fluctuations on either side of the average brightness are transmitted, and this axis must of necessity be the zero-voltage axis in all instances because no d-c component can get through the amplifier.

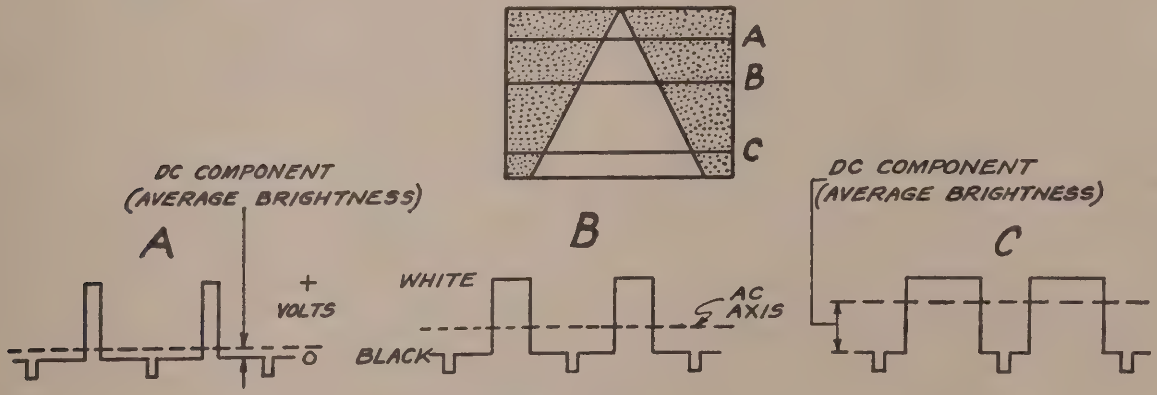


FIG. 7-15

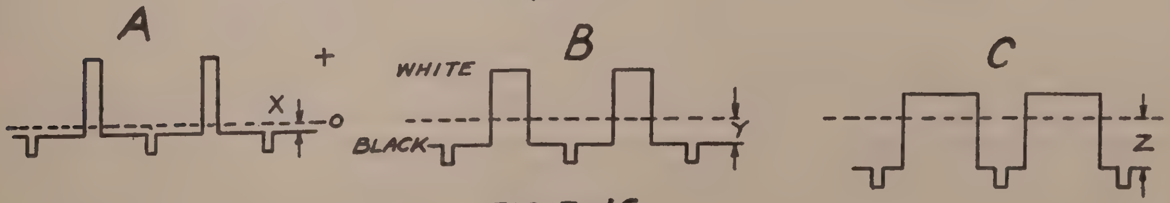


FIG. 7-16

FIG. 7-15.—The video wave produced when two lines are scanned at *A*, *B*, and *C* of the white triangle on the black background that is shown above the waveforms. Since the black level is the same for all three cases, the signal contains the d-c component.

FIG. 7-16.—These are the waveforms of the three signals of Fig. 7-15 after passing through an a-c amplifier. Note that the d-c component has been lost so that black in the signal no longer corresponds to a fixed voltage level as it does in Fig. 7-15.

Having investigated the video signal both with and without the d-c component, let us now examine the manner in which the picture tube is affected by the presence or absence of the d-c component. In Fig. 7-15, we see that black in every case corresponds to the same definite voltage value and this is also true for white and every intermediate shade between black and white. Thus when the voltage of Fig. 7-15 is applied to the grid of the picture tube, the picture will be reproduced without any distortion.

This, however, is not true of the signal in Fig. 7-16, from which the d-c component has been removed. Black no longer corresponds to the same value in all instances, but instead the signal voltage associated with black takes on a value which is entirely dependent upon the average brightness of the strip being scanned. In the same way, this figure shows that white, and every intermediate shade as well, has a different voltage value which is also dependent upon the average brightness of the strip being scanned. A little reflection will show that this gives rise to serious distortion because the same light value in different parts of the picture does not correspond to the same voltage value in the signal. Thus, for example, the grid of the picture tube receives a *different voltage* for the *same shade* of gray in (A), (B), and (C) of Fig. 7-16 and as a result this shade is reproduced differently in the three instances.

In order to remove this source of distortion, it is apparent that the average brightness or the d-c component must be restored. At first glance this seems impossible, for how can the d-c component be restored after it has once been lost in the video amplifier? As a matter of fact, it is generally not possible to restore the d-c component of a pulsating wave after the d-c component has been lost in transmission, because no information relative to the d-c component is contained in the a-c component; the two are entirely independent of each other. Fortunately, however, the sync pulse is transmitted at the end of each line, *and the pedestal on which the sync pulse stands corresponds to a definite black reference level.* These facts make it possible to restore the lost d-c component. Thus to restore the d-c component to the signal of Fig. 7-16 it is only necessary to modify the signal so that all the synchronizing pulses are lined up. When this is done, the signal in Fig. 7-16 is exactly the same as that in Fig. 7-15 and the d-c component has been completely restored.

The solution of the problem depends upon finding this varying d-c voltage and adding it to the signal. This is accomplished in a very simple manner by using a diode to rectify the "black" half of the a-c video signal (which contains the sync signal) and in this way the required voltage is produced. Thus in Fig. 7-16 at (A) this rectification produces the voltage x; at (B) it produces the voltage y; and at (C) it produces the voltage z. In general, the addition of this varying d-c voltage lines up all the pedestals to produce the original signal (Fig. 7-15) with the d-c component restored.

Basic D-C Restorer Circuit

In Fig. 7-17 is shown a straightforward circuit which is used to restore the d-c component in the video signal before the signal is applied to the control grid of the picture tube. The video signal is developed across the 3000-ohm plate resistor *R*1 and fed to the control grid through a 0.1 μf coupling capacitor *C*1. As the sketch shows, the signal has the required positive polarity both at the plate of the video amplifier and the grid of the picture tube. However, the d-c component is not present at the plate side of *C*1, whereas the diode circuit shown in heavy outline has restored the d-c component on the picture-tube grid side.

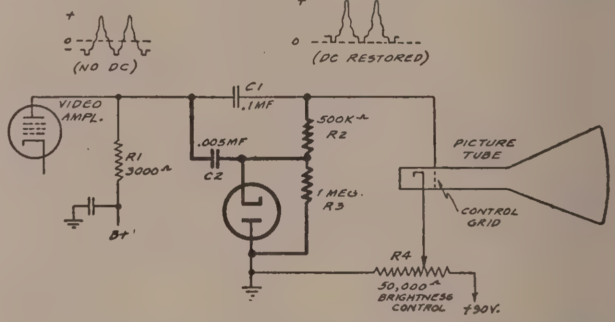


FIG. 7-17.—The heavy portion of the circuit shows a typical d-c restorer circuit arranged to restore the d-c component to the video signal before it is applied to the control grid of the picture tube.

Let us see just how the diode circuit restores the d-c component. Considering the series circuit composed of the plate-load resistor *R*1, the 0.005-μf capacitor *C*2, and the diode in shunt with a 1-meg resistor, we can see that the following takes place: On the positive half of the cycle, the cathode of the diode is swung positive with respect to its plate so that no current flows in the diode circuit. On the negative half of the cycle, however, the cathode is swung negative with respect to the diode plate, and current flows in the diode circuit through *R*3. It is this current flow which charges *C*2 and makes the cathode end of R3 positive, and as a result provides the d-c restoring bias.

Note that the d-c voltage across $R3$ satisfies all the conditions required to line up the sync pulses so as to restore the d-c component. Essentially the action in the diode circuit is such that the *negative* part of the video wave is rectified and the diode current charges $C2$ positively to a value equal to the amount by which the sync pulses in (A) are depressed below the a-c axis. Since the grid of the picture tube is returned to this d-c voltage through the 50,000-ohm resistor $R2$, the d-c voltage is added to the video signal at (A) and "raises" the pedestal so that it lies along the zero-voltage axis, as shown at (B).

Now suppose, as in Fig. 7-16 (C), that the average brightness is higher than that shown at (A) in Fig. 7-15. This results in the pedestal being more negative with respect to the a-c axis, so that the diode produces a more positive voltage which again raises the pedestal to the same zero-voltage axis as at (B) in Fig. 7-15. Thus the action is entirely automatic so that all the pedestals are lined up at the grid of the picture tube, *regardless of the value of the average brightness.*

It is important to understand that the voltage produced at the cathode varies constantly throughout the scanning and that its value at any time depends upon the average brightness of the *portion* of the picture being scanned at that particular time. The time constant (RC) of the diode circuit is designed so that $C2$ will not discharge appreciably during the interval between successive sync impulses. At the same time, the time constant is sufficiently small so that when the average brightness changes, the capacitor is able to change its charge rapidly enough to respond to the new conditions.

Insofar as the grid of the picture tube is concerned, it receives the a-c component of the video signal through $C1$, and *only* the d-c component through $R2$. Although the a-c component is also present at the cathode of the diode, $R2$ acts as a filter resistor to prevent that portion of the video signal present at the cathode from reaching the picture-tube grid through $R2$.

Brightness Control

We have just seen how the d-c restorer circuit automatically lines up all the sync pulses so they are at the same voltage level. For correct operation of the picture tube, the bias on the picture tube must be so adjusted that these aligned pedestals occur at the cutoff or black level. The sync pulses will then lie in the blacker-than-black region and the various shades of gray and white will be reproduced correctly.

Fig. 7-18 shows the illumination characteristic of a picture tube and the manner in which the brightness of the scanning spot depends upon the bias voltage on the control grid. Referring again to Fig. 7-17, it will be observed that the cathode of the picture tube is returned to a potentiometer which makes it possible to vary the bias voltage from 0 to 90 volts.

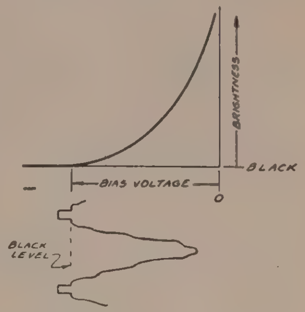


Fig. 7-18.—The effect of bias voltage on the brightness of the image. The brightness control should be adjusted so that the black level on the signal occurs approximately at the cutoff of the picture tube characteristic.

When the bias is highly negative, the tube is cut off and the spot intensity is zero. As the bias becomes more positive, the intensity of the spot increases. For correct operation, the brightness control should be adjusted manually so that the pedestals occur at the black or cutoff points. Once the brightness control has been set, the d-c restorer circuit automatically keeps the pedestals in alignment so that no further adjustment is required.

Grid Leak — Capacitor Restorer

Another type of d-c restorer circuit which is very widely used is shown in Fig. 7-19. In this circuit the output video amplifier tube is operated at zero bias and the grid-cathode elements are used as a diode to insert the missing d-c component. To prevent the loss of the newly restored d-c component which would occur if capacitor coupling were used in coupling the plate to the picture tube, the plate is coupled *directly* to the picture-tube grid.

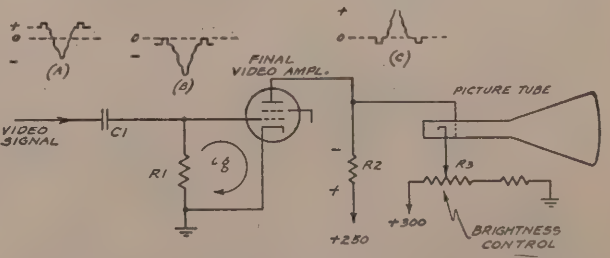


Fig. 7-19.—A widely used type of d-c restorer circuit in which the video amplifier output tube is operated at zero bias so that the grid-cathode elements function as a diode to reinsert the d-c component.

From the explanation already given of the process of d-c restoration, it is easy to understand how circuits of this type operate. The waveform of the video signal on the left side of *C*1 is shown at (A); at this point the signal has a negative polarity and of course the d-c component is missing. When this a-c signal is impressed on the grid of the output tube through *C*1, the positive parts of the video cycle (including the sync pulses) place the grid positive with respect to its cathode so that grid current flows through *R*1. The direction of this grid-current flow is such that the grid becomes negative with respect to ground.

When the average brightness of the signal is small, the a-c axis of the signal will be near the black level, the positive peaks will be small in amplitude, only a small value of grid current will flow, and consequently the grid will have only a slight negative potential on it. On the other hand, when the average brightness is great, the a-c axis is away from the black side, the positive peaks will be large in amplitude, a relatively large flow of grid current will take place, and the grid will be made highly negative. *In each instance, the grid will be made more negative by an amount equal to the height of the sync pulses above the a-c axis of the wave.* As a result of this action, all of the pedestals in the signal are depressed by an amount sufficient to align them to the same zero-voltage level at the grid. This is shown by the signal waveform at (B). Essentially the action here is the same as that previously described for the circuit using a separate diode. The capacitor *C*1 performs the same function of storing the grid charge as in the previous method.

At the plate of the output tube, the polarity of the video signal is reversed and as (C) shows, the signal at this point has the required positive polarity. Since no blocking capacitors are interposed between the plate and the grid of the picture tube, the sync pulses remain in alignment.

As in the previous circuit, provision must be made for initial adjustment of the bias of the picture tube so that the pedestals of the signal will occur at cutoff (black level) on the picture-tube characteristic. To accomplish this, the cathode of the picture tube is returned to a bleeder-potentiometer which makes it possible to place up to a maximum of +300 volts on the cathode. Since the grid can at most have a potential of +250 volts, this makes a bias of —50 volts available on the grid of the picture tube. Under actual operating conditions, the voltage drop across *R*2 makes the grid more negative; this is compensated for by making the cathode of the picture tube less positive by adjusting the brightness control *R*3.

When the receiver is first turned on, there is no voltage drop across the plate-load resistor *R*2 until the output video tube warms up and draws plate current. The final adjustment of the brightness control therefore should not be attempted until the receiver has been turned on for a few minutes. In particular the brightness control should be all the way to the left (with the rotor at +300) until the receiver has warmed up; this is done to avoid possible damage to the picture tube because of excessive beam current.

Video Output of Belmont Model 21A21

The video output circuit of the Belmont 21A21 television receiver, as shown in Fig. 7-20, is an example of the grid leak-capacitor type of d-c restorer. The d-c restorer in this circuit is the same as the basic circuit shown in Fig. 7-19. The negative bias on the grid of the 6AH6 will be determined by the height of the sync pulses above the

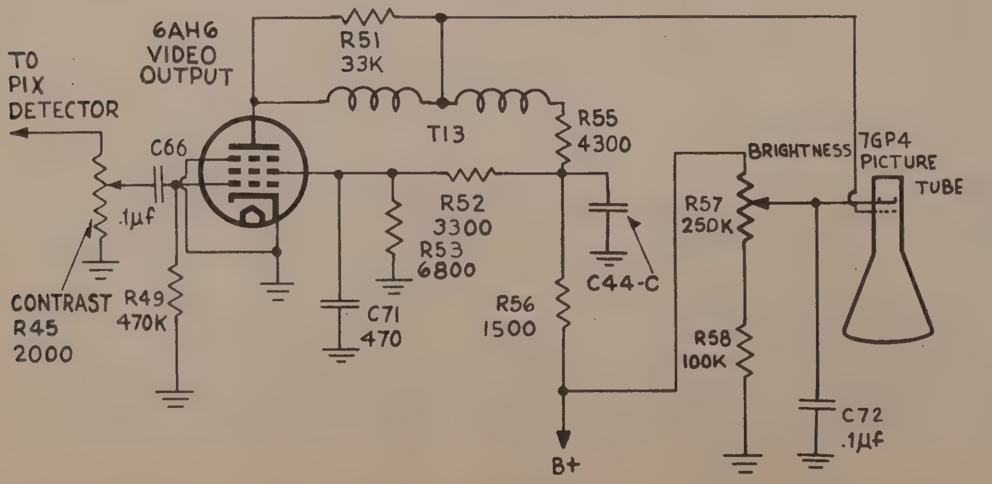


Fig. 7-20.—The video output circuit of the Belmont model 21A21 is an example of the grid leak-capacitor type of d-c restorer, which is the same as the basic circuit shown in Fig. 7-19.

a-c axis. This will line up all the sync pulses at zero level and restore the d.c. exactly as explained with regards to the basic circuit in Fig. 7-19. The time constant of the *C*66-*R*49 combination is equal to 470,000 $\times$ 10^{-7} or 47,000 microseconds. The time between successive horizontal sync pulses is equal to the reciprocal of the horizontal sweep frequency or 1/15,750 which is equal to 63.5 microseconds. It can be seen that the time constant used in this circuit is very much greater than the time between sync pulses. Capactor *C*66 will not be able to discharge between sync pulses and, therefore, the circuit will not respond to changes in brightness of each individual line. Let us now consider the effect of a change in average brightness of the entire picture. This will not take place more often than, let us say, twice a second. The time between changes will therefore be 0.5 second or 500,000 microseconds. This is more than 10 times the time constant of the *RC* circuit used for d-c restoration. Capacitor *C*66 will be able to discharge appreciably in this time and therefore the circuit will respond to changes in average brightness of the entire picture.

The 2000-ohm potentiometer in the input circuit to the 6AH6 varies the signal applied to the grid of this tube. This enables the range of brightness or the contrast of the picture to be adjusted. The potentiometer *R*57 varies the brightness of the picture by changing the bias on the picture tube. When properly adjusted, the scanning spot will be cut off by the blanking signal. Notice that only one video amplifier is employed in this model, as the signal output of this amplifier is sufficient to properly drive the 7GP4. The compensating circuits used in this set are similar to those that have been discussed. Combination shunt and series peaking is used. The coil assembly *T*13 contains both the shunt and series peaking coils.

Video Output of RCA Model 630TS

The video output stage of the RCA model 630TS is shown in Fig. 7-21 as an example of a diode d-c restorer. Comparing this circuit to the basic circuit as shown in Fig. 7-17, we see that it is basically the same circuit but that several new features have been added. First, notice the brightness control. Instead of supplying a d-c bias to the cathode of the picture tube as in the basic circuit, this brightness control supplies a negative bias to the grid of the picture tube. The object is the same in both cases, that is, to set the operating point of the Kinescope at the correct value. In Fig. 7-21 the brightness control *R*152 is adjusted for the correct operating bias of the 10BP4. This is about —50 volts. This voltage is applied to the plate of the diode (1/2 of a 6AL5) through resistors *R*151 and *R*150. The cathode of the diode is connected to the top of the plate-load resistor of the 6K6, *R*147, through the series combination of *R*146 and *C*142, thereby applying the video signal output of the 6K6 to the cathode of the diode. The only time the diode can conduct is when the signal on the cathode goes more negative than the —50 volts or so that is applied to the plate of the diode; this is the same as saying that the plate goes positive with respect to the cathode. If the circuit is adjusted properly this will happen only at the bottom of the sync pulses. The sync pulses at the output of the 6K6 are in the negative direction in order to get the correct polarity on the grid of the Kinescope. When the diode conducts, the capacitor *C*142 will charge and produce a d-c voltage on the cathode of the diode. This voltage is applied to the grid of the Kinescope through the resistor *R*148. This voltage is added to the video signal and will raise the pedestal above a reference level which is equal to the negative voltage applied to the plate of the diode by the brightness control. Thus all the pedestals will be lined up at the grid of the picture tube at the correct operating point of the tube.

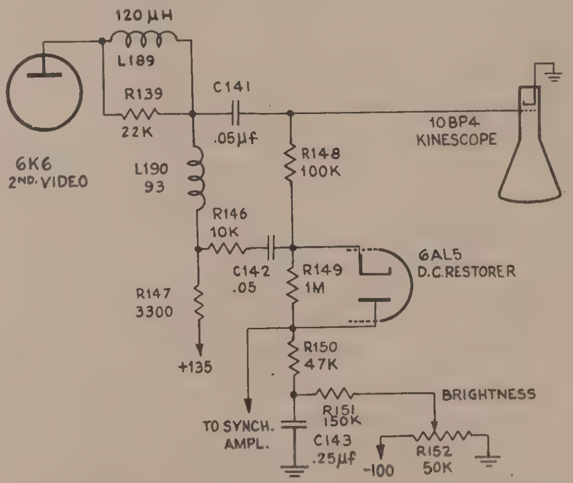


FIG. 7-21.—The video output stage of the RCA model 630TS is an example of a diode d-c restorer. It will be noticed that several changes have been made when it is compared with the basic circuit of Fig. 7-17.

The time constant of the d-c restorer circuit depends mainly on the combination of *C*142 and *R*149 and is equal to 0.05 second. This time constant is sufficiently larger than the time between horizontal sync pulses so that the circuit will respond only to changes in the average brightness of the entire picture. The action is similar to that discussed in the grid leak-capacitor type of d-c restorer. Resistor *R*146 in series with capacitor *C*142 isolates the stray capacitance of the diode circuit from the output circuit of the 6K6. Resistor *R*150 serves as a load resistor for the sync output circuit. Since current flows through this resistor only when the sync pulse occurs and the diode conducts, the sync pulses alone will appear across *R*150 and can be used to supply the input of the sync amplifier. Therefore this circuit performs another function, that of removing the sync signals from the video signal.

The Contrast Control

The "contrast control" is the gain control which determines the magnitude of the video signal applied to the grid of the picture tube. Its counterpart in a sound receiver is the volume control, and in the same way that the sound volume control determines the range of intensities between the loudest sound and the softest sound, so the video contrast control determines the range of light intensities between the highlights and the shadows of the picture. Referring to Fig. 7-18, the setting of the contrast control determines how much of the picture-tube characteristic is used and how bright will be the brightest element of the scene. When the contrast control is not advanced far enough, the picture lacks brilliance and the highlights are comparatively dark; when the contrast control is advanced too far, the picture becomes blurred, there is a loss of detail in the highlights, and in general the intermediate shades are lost.

CHAPTER 8

SYNCHRONIZING CIRCUITS

By Richard F. Koch

The circuits devoted to synchronization in a television receiver are those which remove the sync information contained in the complete video signal and utilize it to control the timing of the horizontal and vertical deflection oscillators. The need for this has been discussed in connection with the subject of the composite transmitted signal. There it was shown that an important part of the transmitted signal is the sync signals, which lie in the "blacker-than-black" region. Without these sync signals, the picture on the cathode-ray tube of the receiver would not stay put; at the very least its motion would be annoying, and in most cases the motion would be so great as to make the presentation useless as entertainment.

We have also seen that the television transmitter broadcasts two different kinds of sync signals—horizontal and vertical. The sync circuits are required to separate these two signals, in addition to the other functions they must perform. Having separated these signals, the sync circuits must route them to the appropriate deflection circuits. We shall discuss the manner in which this is accomplished shortly, but first let us consider some generalities in the characteristics of sync circuits and the restrictions that are imposed upon them by outside influences such as static.

General Requirements

It must be remembered that of all the broadcasting services, television is the most sensitive to static interference. It has been stated (although this statement has been disputed) that a 500-microvolt signal is required to get a good picture on the cathode-ray tube of a television receiver. There are several reasons why such a strong signal is needed. Most important is the difference between the human eye and the human ear, for the former is far more sensitive to visible distortion than the latter is to audible distortion. As a result, flaws in the sound coming from a loudspeaker due to the presence of a certain amount of static may be negligible, but when the same amount of static acts upon a television receiver, the picture on the cathode-ray tube may be intolerably distorted.

One of the weakest points in a television receiver, as regards its sensitivity to static, is the sync system. This is particularly true of the horizontal synchronization, because of the high frequency of the horizontal sweep and the narrow sync pulses employed. (Horizontal sync pulses arrive at the rate of 15,750 per second and are only 4½ to 5½ microseconds wide; vertical sync pulses, on the other hand, arrive in groups occurring only 60 times per second, and these groups are some 190 microseconds in width.) A burst of static may, therefore, mask a substantial number of successive horizontal sync pulses; or, if static occurs in pulse-like form, it may cause the horizontal sweep to sync with it instead of with the broadcast sync pulses. The same burst of static would have much less effect on the vertical sync; in fact, it might easily come between two consecutive groups of vertical sync pulses and have no effect whatsoever on the vertical deflection circuit. If there is a considerable amount of static, some of it will probably arrive at the receiver at the same time as a group of vertical sync pulses, and so affect the operation of the vertical deflection circuit. However, for a given intensity of static, the horizontal deflection circuit is more likely to be affected than the vertical sweep.

Although this difference in susceptibility to interference must be borne in mind, it must also be remembered that it makes little difference to the audience whether the picture is given to jumping vertically or horizontally. In either case they will be annoyed, and so both sync circuits must be protected from interference. The best way of doing this is to keep the signal-to-noise ratio as high as possible. This is accomplished partly by getting the strongest possible signal to the input terminals of the receiver and partly by using circuits that will respond to sync pulses in the desired manner, while effectively rejecting interference such as static. This rejection may be accomplished in one of two ways or by combining the two methods. One way is to use limiters or clippers that prevent any signals corresponding to substantially more than 100-percent modulation from passing through the sync circuits to affect the sweep circuits. Since the sync pulses correspond to the top 20 to 25 percent of the video modulation imposed upon the video carrier at the television transmitter, this limiting action will pass the sync pulses, while cutting off interference riding above them.

Sync Methods: Instantaneous Locking vs. AFC

As stated above, there are two methods of reducing the effects of interference on the operation of the sync circuits. One of these has already been mentioned; namely, clipping or limiting. The other involves an averaging action, causing the sweep to be synchronized at the *average* frequency of the sync pulses. The average frequency of the sync pulses generated at the transmitter is of course constant, and their spacing is quite regular. However, extraneous effects, such as static, may modify the *apparent* spacing of these pulses.

A way in which this may occur is shown in Fig. 8-1. The three pulses are shown here in a very simple manner. Simultaneously with one of these pulses a burst of static occurs. As a result, the apparent amplitude of the concurrent pulse, as seen by the receiver sync circuit, is much greater than its fellows, *and,* what is more important, it seems to start sooner. Looking at the portion of the figure illustrating the received pulses, you will observe a line labelled "required sync pulse height." This is the height to which the sync pulses must rise in order to perform normal synchronizing action. Since simple synchronizing is something like the action of an on-off switch, a pulse failing to reach the required level will have the effect of no pulse at all, while a very high pulse will affect the circuit pretty much like one of the proper (lower) height. Since the middle pulse, riding on top of the static burst, reaches the height required for synchronizing sooner than it would in the absence of the static, it *seems* to be spaced nearer to the preceding pulse than it was when it was transmitted. Similarly, the apparent spacing between the beginnings of the second and third pulses *seems* to be greater than the true spacing. If there is appreciable static, this effect may be repeated frequently, causing considerable jitter in the picture seen on the picture tube.

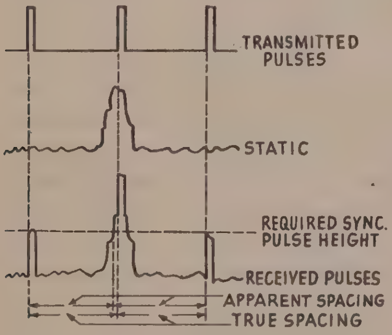


Fig. 8-1. — Apparent change in sync pulse spacing caused by static.

As was pointed out previously, the horizontal sweep circuit is more susceptible to synchronizing difficulties of the sort illustrated by Fig. 8-1 than is the vertical sweep circuit. For this reason it has seemed desirable in some television receivers to use an afc (automatic frequency control) circuit to provide horizontal sync. In this method of controlling the horizontal sweep, the individual sync pulses do not exert a direct effect on the sweep oscillator. Instead, the sweep frequency is locked to a value corresponding to the *average* spacing of perhaps one hundred or so sync pulses. (How this is accomplished will be seen later, when we consider the horizontal sync circuit of the RCA television receiver model 630TS.)

Now let us re-examine Fig. 8-1 in order to see the benefit of this averaging process. In this simplified example, the apparent spacing between the first and second pulses is less than the true spacing with which these pulses left the transmitter, and the apparent spacing between the second and third pulses is greater than the true spacing. However, the spacing between the first and the third pulses as they are received is the same as when they were transmitted, because the static has had very little effect upon them. Since the first and third pulses have not experienced an *apparent* shift in position, as the second has, the *average* spacing of the three pulses is equal to the *true* spacing. This is so because the spacing from the first pulse to the second has apparently decreased, while the spacing from the second to the third has apparently increased by the *same* amount as the spacing from the first to the second decreased. The decrease and the increase balance each other, and thus the *average* spacing of the three pulses is equal to their *true* spacing.

We pointed out before that when the *individual* sync pulses are used to synchronize the sweep, static may cause the picture on the cathode-ray tube to jitter. This occurs because the apparent spacing of the pulses is affected by the static. In the example of Fig. 8-1 we have seen that the static does not affect the average spacing of the three pulses shown. Therefore, if the average value of the sync pulse spacing is used to determine the sweep frequency, static will not cause the picture to jitter. (The example of Fig. 8-1 is simplified to the point where it is not entirely realistic, for under actual operating conditions static would probably affect a series of pulses, rather than just one out of three. However, if we considered a much larger number of pulses that three, we would find that in actual practice the averaging effect holds, just as in the simplified example.)

Sync Separation

The circuits used to separate the sync pulses from the picture portion of the signal are called "sync separators" or "clippers." These circuits are designed so that they are responsive only to the sync pulses, which lie above the pedestals (that is, in the "blacker-than-black" region); thus they reject the picture portion of the signal. This rejection is necessary to prevent the video signal from traveling through the sync circuits to the sweep oscillators. The video signal might affect the operation of the sweep oscillators, causing them to fall out of their required synchronism. The clipper or sync separator is generally followed by additional amplification, and finally the sync signal is fed to a selecting circuit which separates the horizontal sync pulses from the vertical sync pulses. This separation is accomplished by circuits which depend for their action upon the difference in the time duration of the two types of pulses.

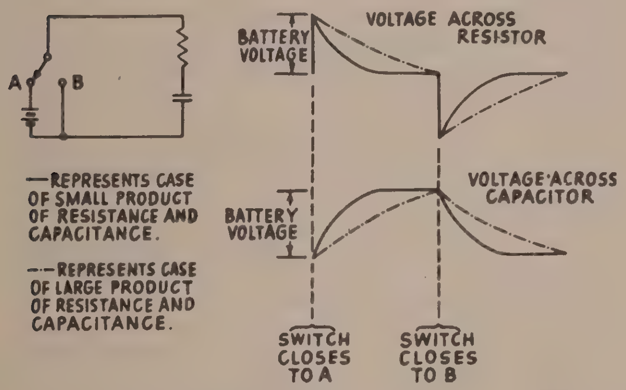

A
B
BATTERY VOLTAGE
VOLTAGE ACROSS RESISTOR
VOLTAGE ACROSS CAPACITOR
BATTERY VOLTAGE
—REPRESENTS CASE OF SMALL PRODUCT OF RESISTANCE AND CAPACITANCE.
-·-REPRESENTS CASE OF LARGE PRODUCT OF RESISTANCE AND CAPACITANCE.
SWITCH CLOSES TO A
SWITCH CLOSES TO B

Fig. 8-2, above.—Sync separator circuit consisting of a resistor and capacitor in series. The voltages that appear across each when the capacitor is charged from a battery and then discharged, in both cases through the resistor, are shown at the right.

The circuits which discriminate between long (vertical) and short (horizontal) sync pulses are simple combinations of resistors and capacitors. Fig. 8-2 shows a resistor and capacitor in series and the voltages that appear across each when the capacitor is charged from a battery and then discharged, in both cases through the resistor. Two sets of curves have been drawn, corresponding to two combinations of resistance and capacitance. It may be seen from this illustration that when the product of resistance and capacitance is large, a relatively long time is required to charge the capacitor to full voltage or to discharge it to zero. However, when this product is small, a relatively short time is required. Because of the relationship between the time required for charge or discharge and the product of resistance and capacitance, this product is called the "time constant" of the circuit.

Now let us look at Fig. 8-3, in which is shown the behavior of series resistance-capacitance circuits such as those shown in Fig. 8-2, when series of pulses are impressed across them. In this figure two special cases are illustrated; these cases are very important in television receivers, as will be seen later. Part (A) shows the signal observed across the resistance when narrow pulses are impressed across a series combination having a short time constant. (In this discussion it must be remembered that the terms "short" and "long" are relative.)

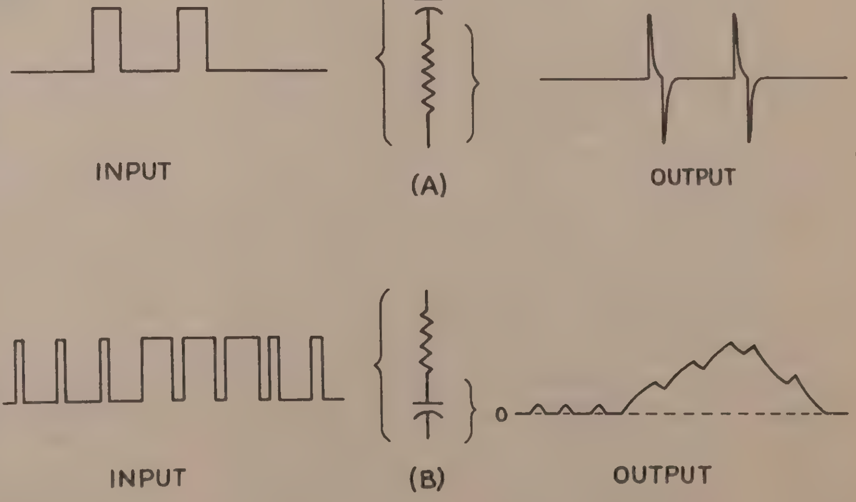


Fig. 8-3, right.—Effect of series resistance-capacitance circuit on input pulses, showing (A) the signal observed when like narrow pulses are applied and (B) when the input consists of two different kinds of pulses.

Comparing this with the resistor voltage curve for the short time constant case in Fig. 8-2, we see that in both cases the voltage rises immediately to the peak value of the applied voltage, and then drops quickly back to zero. When the applied voltage drops to zero and the capacitor discharges, the flow of current is opposite to that which existed during the charging pulse, and a pulse similar to the charging pulse, but of opposite polarity is produced. The output then remains at zero until the next input pulse arrives. It may also be observed that there is a very short interval between the pulses of opposite polarity when the output is at zero. Each pair of output pulses corresponds to one input pulse.

In Fig. 8-3(B) a somewhat more complicated case is illustrated. Here the input consists of two different kinds of pulses. The output in this case is obtained across the capacitor, and the time constant of the circuit is long. Because the time constant is long, the first narrow pulse produces a small rise in voltage in the output; this rise continues for the duration of the pulse, and at its conclusion the output voltage returns to zero. Compare this with the capacitor charge and discharge curves in Fig. 8-2, for the long time constant case. Returning to Fig. 8-3(B), we see that the second and third narrow input pulses produce the same kind of low, triangular pulses in the output as that produced by the first pulse. However, the fourth pulse is of relatively long duration and the charge on the capacitor—and, therefore, the output voltage—rises to a much higher level than during the short pulses. When the pulse ends, the charge starts to leak off at a rate determined by the resistance; but the next pulse, also a wide one, comes along very soon, and the drop in output voltage is only a small part of the rise that occurred during the first wide pulse. The new rise in charge—and output voltage—that takes place during the application of the second wide pulse to the circuit therefore starts from a level higher than that at which the rise produced by the first wide pulse started. As a result, the output voltage rises to a new high value at the end of the second wide pulse. During the interval between the second and third wide pulses there is again a drop in voltage; nevertheless, the point from which the rise due to the third wide pulse starts is higher than the starting points of any of the previous rises. The end of this third pulse therefore sees the output voltage at the highest value yet. A drop, of course, follows and then another rise due to the succeeding narrow pulse. However, because the duration of this narrow pulse is less than that of the space that preceded it, the rise that it produces is less than the drop which it follows. Therefore, at the end of this narrow pulse the output voltage is lower than it was at the end of the third wide pulse. The interval before the occurrence of the next narrow pulse is relatively long, and so the output voltage falls appreciably during this time. Then, during the narrow pulse it rises slightly again, and finally falls to zero in the interval succeeding the last narrow pulse.

Let us review briefly the actions illustrated in Fig. 8-3(B). The narrow pulses produce small rises in voltage, small because the capacitor is able to acquire only a small charge during the short interval of each pulse in this long time constant circuit. In the long intervals following the narrow pulses all the charge acquired by the capacitor during the preceding pulse leaks off. When the pulse-and-space situation is reversed, however, the output voltage changes materially. Here, the pulses are wide and the intervals between them are narrow, so that the capacitor has an opportunity to acquire a rather high charge when a pulse is applied, and only a small part of this charge leaks off in the interval between pulses. Therefore, when the wide, closely-spaced pulses are applied, the output voltage rises to a point considerably above zero; while when narrow, widely-spaced pulses are the input, the output voltage is very low. After the three wide pulses, narrow pulses are applied again, and the output voltage drops back to zero. Of course, if narrow pulses continue to be applied, they will produce low triangular pulses in the output, just as the first three did, but for most of the time the output will be zero.

Looking once more at Fig. 8-3(B), we observe a very important interrelation between the input and output waves: when the input consists of narrow pulses, spaced far apart, the output consists of low triangular pulses, and the general level of the output is practically zero. However, when wide, closely spaced pulses are applied to the input, the output level rises considerably, without any fall to zero between input pulses. Finally, when narrow, far-spaced pulses are reapplied, the output falls nearly to zero, and conditions return to those that existed when the narrow pulses were applied originally. Thus the circuit discriminates between narrow and wide pulses by producing a small, almost insignificant output for the former and a large output for the latter.

To illustrate the principles involved in sync circuits, we shall consider some specific examples from television receivers on the market today. As an example of sync clipping and separation we shall consider the circuits used in the Belmont models 22A21-22AX21 and 22AX22. RCA model 630TS has been chosen as an example of the afc type of horizontal sync control.

Belmont Models 22A21, 22AX21 and 22AX22

As the block diagram in Fig. 8-4 shows, the video signal is taken from the first video amplifier and passes through the sync clipper, which removes most of the video signal from the sync signal. The sync signal is then fed to a sync limiter stage which fixes the height of the sync signals, removing small variations in amplitude that may persist after the action of the avc in the r-f and i-f amplifiers. The output of the limiter is applied to two pulse selectors—one for horizontal and one for vertical sync pulses. These selectors pass the sync pulses required by the multivibrator sweep oscillators to which they are connected, while virtually eliminating the undesired pulses.

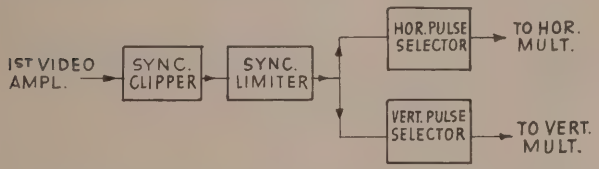


FIG. 8-4.—Block diagrams of the sync circuits used in the Belmont models 22A21, 22AX21, and 22AX22.

On the whole, the block diagram presents quite an imposing array of circuits, considering that it represents but a small part of all the circuits in the receiver. However, synchronization is so important in the operation of the television system that great care must be taken to make synchronization positive in action,

Referring to the partial schematic of Fig. 8-5, we see that the video signal is fed to the grid of the sync clipper tube, one half of a duo-triode type 6SL7-GT. The distinguishing features of this stage are that the tube is operated at the very low plate voltage of 5.6 volts, and at a bias close to cutoff, provided by the flow of grid current through the large grid resistor *R*60. (The very low plate voltage is attained by using a very low effective supply voltage—about 8 volts—obtained from a bypassed bleeder across the 300-volt plate supply.) As the waveform at the grid of this tube shows (Fig. 8-5), the sync pulses are the most positive part of the input signal. Since the plate voltage of this tube is very low, only a very small negative signal is required to drive it into cutoff, particularly since there is already a relatively high (considering the low plate voltage) negative bias due to *R*60. Therefore the tube is cut off by the picture components of the video signal, and passes plate current only during the most positive part of the signal, that is, the part devoted to the sync pulses. As a result, the picture components of the signal have no effect on the plate current (except to keep it at zero), but the sync pulses cause it to flow in corresponding pulses. These pulses of current produce voltage pulses across *R*62 (and *R*64), as shown in Fig. 8-5. The signal has thus been transformed in a desirable manner by being passed through the sync clipper; the picture components have been eliminated, so that they cannot cause synchronization at the wrong time, while the sync pulses have been retained.

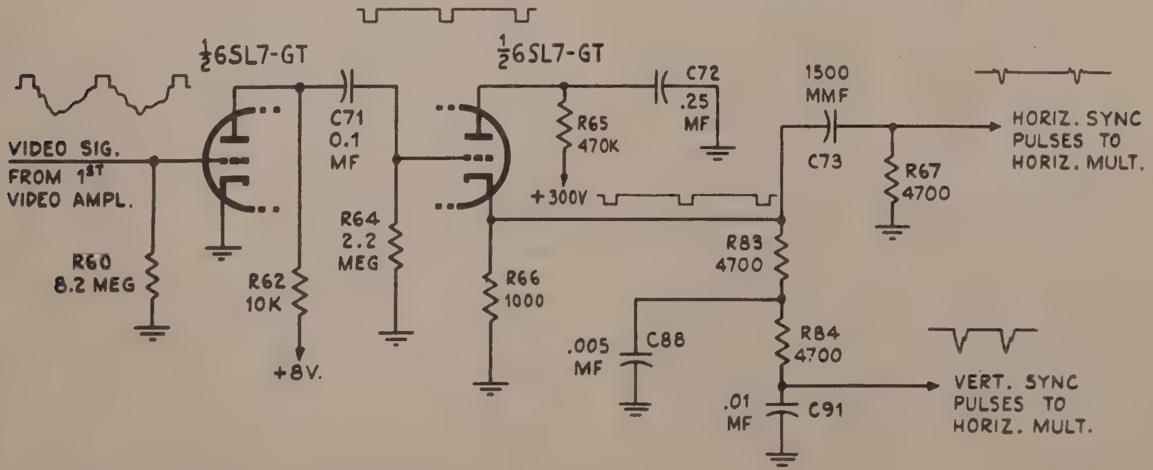


FIG. 8-5.—Partial schematic of the sync circuits of Belmont models 22A21, 22AX21, and 22AX22.

The clipped sync signal is next fed to the grid of the second half of the 6SL7-GT, which is a cathode follower providing limiting action. The combination of *R*65 and *C*72 has a time constant of nearly one-eighth of a second, which is sufficiently long compared to the intervals between sync pulses to keep the plate of the tube at a virtually constant voltage. This plate voltage is 63 volts, and the cathode voltage between input pulses is about 0.5 volt. Since the input pulses are negative and of an amplitude almost equal to the plate voltage of the sync clipper stage (in other words, about 5 volts), they will cut off the limiter tube. Thus, the limiter acts as a switch, passing plate current in the absence of pulses and stopping it in their presence. As a result, the amplitude of the pulses appearing at the cathode of the sync limiter stage is unaffected by the amplitude of input pulses, but the timing of the input and output pulses is exactly the same.

Connected to the cathode of the sync limiter are the two pulse selector networks, which separate the horizontal and vertical sync pulses. The combination of *C*73 and *R*67 has a time constant (about 7 microseconds) which is approximately equal to the width of a horizontal sync pulse. These pulses are, therefore, passed through it with little change, as shown by the final output to the horizontal multivibrator in Fig. 8-5. When individual vertical sync pulses reach this circuit they are converted to sharp double pulses, somewhat as shown in Fig. 8-3(A), for their duration is approximately three times the time constant of the combination of *C*73 and *R*67. Because the input pulses are negative, the first pulse of each pair will also be negative, and the second positive. Referring to the standard television signal shown in Fig. 1-21, we see that the beginning of every other vertical sync pulse lies exactly where a horizontal sync pulse ordinarily lies. Therefore during each vertical sync pulse group there will be generated a series of negative pulses, equally spaced, and every other one will correspond exactly in position to the point where a horizontal sync pulse might be found. (There are also generated positive pulses, as previously mentioned, but these have negligible effect on the horizontal multivibrator.)

One might well ask, "What effect will the extra negative pulses in this interval have on the horizontal multivibrator?" The answer is, "None that is significant." The reason for this is that the multivibrator's frequency is only partly dependent on the sync pulses; sync pulses that arrive at or very near the proper point in the multivibrator's operating cycle will affect it, holding it at the desired frequency. However, pulses whose timing is far off will have negligible effect unless their amplitude is very large. In this circuit the sync limiter keeps the pulse level at a point where the extra pulses due to the vertical sync pulses will not interfere with the proper operation of the horizontal multivibrator.

Let us refer once more to Fig. 1-21. Here we see that each vertical sync pulse group is both preceded and followed by equalizing pulses. These pulses are half the width of the horizontal sync pulses and every other one lies where a horizontal sync pulse ordinarily lies. Because they are so narrow they are passed by the horizontal pulse selector just as the horizontal pulses themselves are. Their spacing, and therefore their effect on the horizontal multivibrator, is just the same as that of the negative pulses generated by *C*73 and *R*67 from the vertical sync pulses. Thus the synchronization of the horizontal multivibrator is maintained during the vertical synchronization period, even though horizontal sync pulses as such are not present in this interval. In other words, the vertical sync and equalizing pulses are made to do double duty, for they work in the horizontal sweep circuit as well as the vertical sweep circuit. (This principle of double duty applies to television receivers generally; it depends upon a fundamental characteristic of the standard television signal, and all receivers intended for reception of this signal take advantage of this.)

The complete sync signal is also fed into the vertical pulse selector consisting of *R*83 and *C*88. An additional resistor-capacitor combination (*R*84 and *C*91) is used in order to increase the effectiveness of the vertical pulse selection. The action of *R*83 and *C*88 in differentiating between the vertical and horizontal sync pulses is a direct result of the longer duration of the vertical pulses. Fig. 8-3(B) shows how a circuit of this sort distinguishes between narrow and wide pulses. The output builds up to a maximum at the end of the last wide pulse (in the case of *R*83 and *C*88 the last pulse in each vertical sync pulse group), and then steadies down almost to zero, remaining at this value until the beginning of the next vertical sync pulse group.

To increase the circuit effectiveness, another similar resistor-capacitor combination (*R*84-*C*91) is used in series with the first as shown in Fig. 8-5. This circuit accomplishes a further discrimination between the vertical and horizontal sync pulses so that a sharper output pulse is produced once during each field. The peak of this pulse is used to synchronize the vertical deflection oscillator.

For exact interlacing it is important that the vertical pulses produced at the output of the vertical selector circuit be the same on alternate fields. The equalizing pulses, described in chapter 1, accomplish this by making the conditions exactly the same before and after the transmission of the actual broad vertical sync pulses. As a result capacitors *C*88 and *C*91 charge up to the same values of peak voltage on both the odd and even fields, so that the vertical oscillator is maintained in the correct timing.

AFC Circuits

The block diagram in Fig. 8-6 shows the general arrangement of the sweep afc circuit used in the RCA model 630TS television receiver. This may be recognized as bearing a similarity in principle to the afc schemes used for correcting errors in tuning superheterodynes. In the case of the superheterodyne tuning, it is desired to overcome an error in the frequency of the local oscillator frequency. This error is such that the actual intermediate frequency generated is not that for which the set was designed. In order to produce a correction, part of the i-f signal is fed into a discriminator, whose output depends upon the error in the i.f. This output is a d-c voltage, whose amplitude depends upon the magnitude of the error, and whose polarity depends upon the direction of the error. That is, if the error is due to the oscillator being tuned too high, the output of the discriminator will have, let us say, positive polarity, and then if the oscillator frequency were too low, the discriminator output would be negative. This control voltage (the output of the discriminator) is applied to a reactance (or oscillator control) tube, which is part of the tank circuit of the oscillator. This tube (or, rather, the circuit of which this tube is a part) has the property of acting like a variable reactance, with the reactance dependent upon the control voltage applied to the tube. Since this variable reactance (which may be either capacitive or inductive) is part of the oscillator's tank, it affects the frequency of the oscillator. Therefore, the error signal from the discriminator can be used to change the frequency of the local oscillator, and if the circuit is connected properly, this change will be such as to reduce the error. In this manner a self-correcting system is set up, which tends to correct for errors that may exist in its initial adjustment.

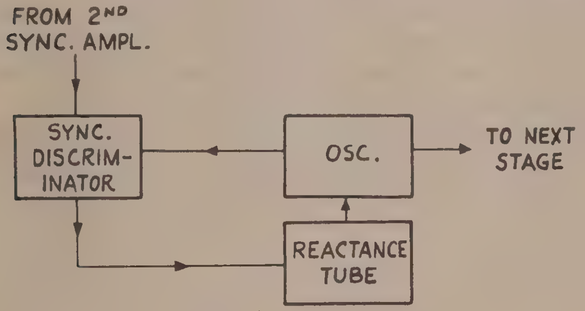


Fig. 8-6.—Block diagram of sweep afc circuit as used in RCA model 630TS. In this circuit the pulses from the second sync amplifier are of primary importance in determining the output from the discriminator.

Fig. 8-7 shows a block diagram of an afc system of the type just described. Let us contrast it with that shown in Fig. 8-6. There are two differences in the inputs to the discriminators shown. One of these is minor. It is this: in Fig. 8-6 the oscillator feeds the discriminator directly, whereas in Fig. 8-7 it does not. However, in the latter case we are not concerned directly with the frequency of the incoming (r-f) signal, and so we assume that any error in the intermediate frequency is due to mistuning of the local oscillator. Therefore, the discriminator input depends directly upon the local oscillator frequency, even though there is not a direct signal path between them.

The other difference between the two figures is a very important one. In Fig. 8-7 there is only *one* input to the discriminator, but in Fig. 8-6 there are *two*. What is the significance of this difference? In a circuit of the type illustrated in the latter figure, the tuning of the discriminator exercises a direct control over the frequency. In the circuit of the former figure, the sync pulses derived from the second sync amplifier are of primary importance in determining the output from the discriminator. In this case the discriminator serves only to compare the frequency of the signal to be controlled with the sync pulses, and its tuning is of secondary importance.

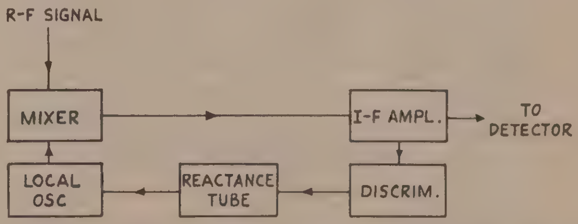


Fig. 8-7.—Block diagram of type of afc scheme used for correcting errors in superheterodyne tuning. In contrast with the circuit of Fig. 8-6, the oscillator does not feed the discriminator directly, and, more important, there is only one input to the discriminator.

A complete schematic diagram of the horizontal sweep oscillator and its associated afc circuit used in the RCA model 630TS receiver is shown in Fig. 8-8. The oscillator tube is *V*125; it operates in a Hartley type circuit. The oscillator coil serves a dual purpose—its basic function and as the primary of the discriminator transformer *T*108. The oscillator coil is tuned by the parallel combination of capacitors *C*169 and *C*164 and the reactance tube *V*124, as shown in Fig. 8-9. In addition to the reactance tube there is another element in the tank circuit that is not a fundamental part of the oscillator. This is the 10-ohm resistor *R*194. Ordinarily the oscillator tank is tuned by a pure capacitance connected from one end of the inductance to the other, but in this case there is a small resistance in series with the capacitance. The total capacitance (*C*169 + *C*164) is 16,200 $\mu\mu f$. At the horizontal sweep frequency of 15,750 cycles the corresponding reactance is approximately 620 ohms. Therefore the 10-ohm resistance will have very little effect upon the current through the capacitive branch of the tank, and the series *RC* combination will behave as though it were almost pure *C*.

What, then, is the function of *R*194? Look again at Figs. 8-8 and 8-9. Reactance tube *V*124 is connected directly across *C*169 and *C*164. (Capacitor *C*173 is simply a d-c blocking capacitor.) In addition, its cathode is connected to the junction of these capacitors and resistor *R*194, and therefore the cathode voltage of the tube is the voltage appearing across *R*194. There are two components of current through *R*194; one is the usual d.c. found in a cathode resistor, the other is the capacitive tank current of the oscillator. As was pointed out above, the effect of *R*194 on the capacitive circuit is so small that the oscillator-frequency current passing through it is almost purely capacitive. To understand the desirability of this capacitive component of current in *R*194, let us briefly consider the manner in which a reactance tube operates.

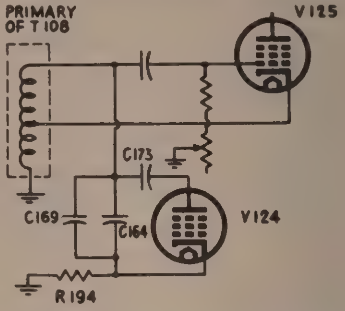

PRIMARY OF T108
V125
C173
C169
C164
V124
R194

Fig. 8-9.—Schematic diagram of the horizontal sweep oscillator circuit of RCA model 630TS.

Reactance Tube Operation

In Fig. 8-10 we see a closed box containing an unknown circuit, and having two terminals. We cannot open the box, but we can measure the current flowing from the box through whatever external circuit may be connected, and we can measure the voltage between the terminals. In addition, we find that the current and voltage are 90° out of phase; this means that the box contains a pure reactance of some sort. Knowing this, we can apply Ohm's law in the form it takes for reactances, namely, namely,

$$X = \frac{E}{I}$$

From this equation we can calculate the size of the reactance contained in the box, and thus we have learned much of the box's contents without

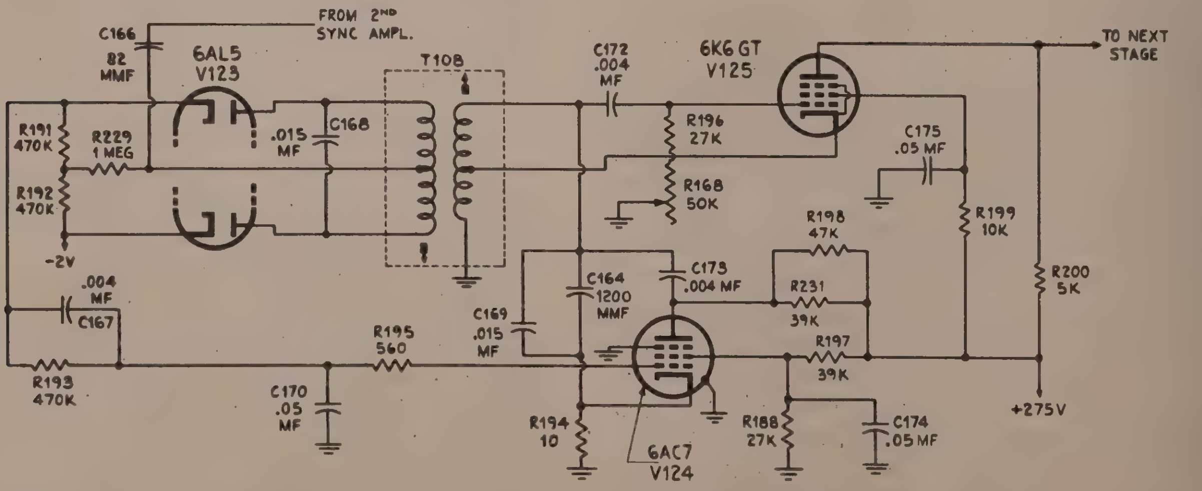

C166
82 MMF
6AL5
V123
FROM 2ND SYNC AMPL.
T108
C172
.004 MF
6K6 GT
V125
TO NEXT STAGE
R191
470K
R229
1 MEG
.015 MF
C168
R196
27K
C175
.05 MF
R192
470K
R168
50K
R198
47K
R199
10K
-2V
.004 MF
C167
C173
.004 MF
C164
1200 MMF
R231
39K
R200
5K
C169
.015 MF
R195
560
R197
39K
R193
470K
C170
.05 MF
R194
10
R188
27K
C174
.05 MF
+275V
6AC7
V124

Fig. 8-8.—Complete schematic diagram of the horizontal sweep oscillator and its associated afc circuit used in the RCA model 630TS.

opening it. Now suppose the box contains a device which can generate a current 90° out of phase with the voltage applied across the terminals of the box. Then, to us outside the box, the device appears to be some sort of reactance. If, further, the amount of current generated can be controlled, the apparent reactance of the device in the box can thereby be controlled. This relationship stems from Ohm's law for reactances, as stated above. For example, if the current is doubled (and the voltage is unchanged), the reactance is halved.

If we can build a device that will actually do what the imaginary device in the box does, we will have a controllable reactance. Such a device can be built using a pentode tube. From an examination of the characteristic curves of a pentode, such as the 6AC7, we see that the plate current is almost independent of plate voltage over a wide range of values. However, the current does depend upon the control grid voltage. Thus it is possible to connect the tube with its plate and cathode across a source of alternating voltage, without having this voltage affect the plate current. However, if an alternating voltage is applied between the grid and cathode, the plate *current* will change in phase with the alternating grid voltage. Now, suppose that the alternating voltages applied in the grid and plate circuits are of the same frequency, but 90° out of phase. Then the plate current will have an alternating component of the same frequency but 90° out of phase with the plate voltage (since it is in phase with the grid voltage). Thus we have constructed a reactance using a pentode, for the current flowing in the plate circuit is 90° out of phase with the voltage across the plate circuit.

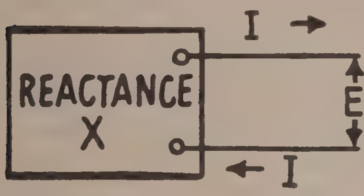


FIG. 8-10. — This closed box having two terminals and containing an unknown circuit illustrates in a simple way how it is possible to obtain a controllable reactance using a pentode tube.

It still remains to be seen how the reactance we have constructed can be made controllable. Look again at the pentode's characteristic curves. One of these shows how the transconductance of the tube varies with the bias on the control grid. The transconductance is a measure of the sensitivity of the plate current to changes in grid voltage; it tells us how much the plate current will change for a change in grid voltage. From this we see that a given change in grid voltage will cause a larger change in plate current when the bias is low than when it is high. Thus the reactance tube we have constructed can be controlled by varying its grid bias.

Sweep AFC Circuit in RCA Model 630TS

Having considered some of the basic principles upon which reactance tube circuits are founded, let us return to our specific example. Fig. 8-11 shows it in simplified form. We have already seen that the current through *R*194 is almost purely capacitive and, therefore, leads the voltage across the tank circuit by almost 90°. Since the current through a resistor is in phase with the voltage across it, the sweep frequency voltage across *R*194 therefore leads the voltage across the entire tank circuit by 90°. Since the full tank voltage is applied to the plate of the tube through *C*173 (whose reactance is very low, and can therefore be neglected), the cathode and plate voltages are 90° out of phase, with the cathode voltage leading. The signal applied to the grid is a d-c control voltage from the discriminator; so far as sweep frequency signals are concerned, the grid appears to be at a fixed potential. The grid bias is approximately two volts; that is, the grid voltage relative to the cathode is approximately —2 volts. In other words, the average cathode voltage is +2 relative to the grid. Then, if the oscillator signal voltage on the cathode at a certain instant is driving the cathode so that it is more positive with respect to ground than its average voltage, this will also make the cathode more positive with respect to the grid than its average value. This is the same as saying that at this instant the grid is becoming more negative with respect to the cathode. The opposite, of course, is also true. When the oscillator signal on the cathode is going negative, its effect is the same as the same signal applied to the grid, but going positive. This can be summed up by saying that a signal applied to the cathode, with the grid voltage fixed, has the same effect as the identical signal, 180° out of phase, applied to the grid with the cathode voltage fixed. Thus, the reactance tube in the RCA model 630TS operates as though the sweep frequency signal were applied to its grid, lagging the voltage applied to the plate by 90°. (It appears to be lagging, when considered in terms of

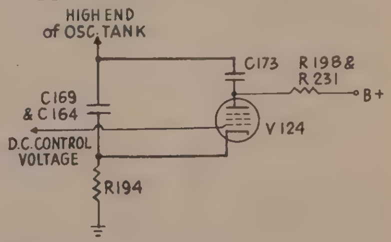


FIG. 8-11.—Simplified circuit of reactance tube used in RCA model 630TS.

its effect at the grid, because of the 180° difference between the grid and cathode as regards inputs to one or the other.)

As was stated above, an alternating voltage applied between grid and cathode will produce an alternating component of plate *current* in phase with the grid signal. In our example, therefore, the plate current lags the plate voltage by 90°. As a result, this reactance tube behaves like an inductance, for in an inductance the current lags the voltage across it. We have seen that when the bias (d.c.) on a tube is increased, the variation (a.c.) in plate current due to a change (a.c.) in grid voltage decreases. Thus, if the bias on the reactance tube is increased, the sweep-frequency component of plate current decreases. Referring once again to Ohm's law for reactances, we recall that a decrease in current is accompanied by an increase in reactance. In an inductive circuit an increase in reactance at constant frequency means an increase in inductance. Finally, then, tube *V*124 acts as an inductance whose value varies with the grid bias, being relatively large for a high bias and small for a low bias.

Returning to Fig. 8-9, we see that the oscillator tank circuit consists of two capacitors (*C*169 and *C*164) and two inductors (the primary of *T*108 and the effective inductance of the reactance tube, *V*124). Since one of the components of the tank—the reactance tube—is a variable controlled by the discriminator, a means is provided for automatically controlling the sweep frequency. The next question is how the discriminator functions to provide the control voltage necessary for the reactance tube.

Fig. 8-8 shows that a coil (the secondary of *T*108) is coupled to the horizontal oscillator tank coil (the primary of *T*108). This coupling provides a sine-wave signal from the oscillator for the diode *V*123. Since the coil is center-tapped, the signals fed to the two diode plates are of opposite phase and equal amplitude. In addition to the 15,750-cycle sine-wave signal from the oscillator, sync pulses are applied to the diodes. These sync pulses come from the second sync amplifier through the short-time-constant circuit *C*166-*R*229. (As explained previously, a short-time-constant circuit is used to separate the horizontal sync pulses from the vertical sync pulses.)

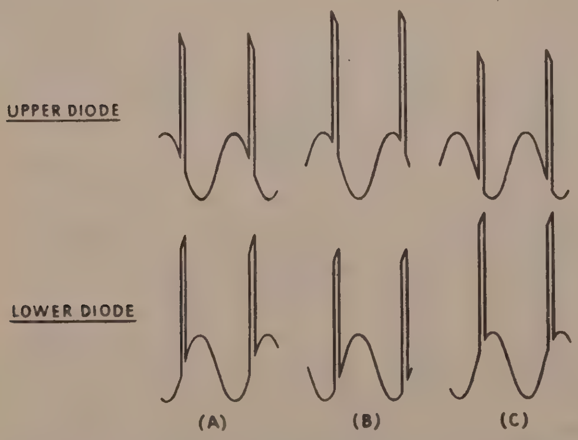


Fig. 8-12.—Combined sync pulse and sine-wave signals at the diode plates when (A) the natural frequency of the horizontal oscillator is correct, (B) the natural frequency is low, and (C) the natural frequency is high. In (B) and (C) the shift of the pulses with respect to the sine waves has been exaggerated to make it more apparent.

Since the sync pulses are applied to the center tap of the secondary of *T*108, they appear at the diode plates in phase, whereas the sine-wave signals are of opposite phase. As a result, the combined signals at the diode plates will not be the same. These signals are shown in Fig. 8-12 for three cases. In part (A) we see the case where the natural frequency of the horizontal oscillator is correct. By this is meant that frequency at which the oscillator would operate if no sync pulses were applied to the discriminator. It is called "natural" because it is the frequency that would exist in the absence of outside influences (sync pulses). In (B) is shown the condition existing when the natural frequency is low, while in (C) is represented the high natural-frequency condition. In the latter two cases it may be seen that the pulses are shifted in phase with respect to the sine wave. This shift has been exaggerated in the figure so as to be more apparent; but although in actual practice only very small shifts occur, it is they that produce the required discriminator action.

In Fig. 8-12(A) the pulses lie on the sine waves at the 0° point for the lower diode and at 180° for the upper diode. Therefore under these conditions the sine waves add nothing to the pulses (the sine of 0° and the sine of 180° are zero), and so the peak-to-peak amplitudes of the plate signals of the two diodes are equal. The voltages produced by rectification of these signals appear across *R*191 and *R*192, which are equal. Since the plate signals are equal, the rectified voltages are also equal. Now the voltages across *R*191 and *R*192 are in opposition, so far as the output to the reactance tube is concerned. This is so because the currents through the individual diodes flow in opposite directions from the junction of the two resistors. Current in the upper diode flows up through *R*191, making the upper end of the resistor positive with respect to the junction. Current in the lower diode flows down through *R*192, making the junction negative with respect to the lower end of *R*192. Thus the junction is negative with respect to the —2v. bias source, but the top of *R*191 is positive with respect to the junction. In this way the voltage across *R*191 opposes that across *R*192. If the voltages are exactly equal, as is the case when the oscillator's natural frequency is correct, they will just balance out, and the control voltage applied to the reactance tube will be the —2v. applied to the lower end of *R*192 from the bias supply.

In the case illustrated by Fig. 8-12(B), the peak-to-peak plate voltage of the upper diode is greater than that of the lower. Therefore, the rectified voltage across *R*191 is greater than that across *R*192. The net effect of this is to make the upper end of *R*191 positive with respect to the lower end of *R*192. As a result, the —2v. bias is partially canceled, and the bias (control voltage) on the reactance tube is lowered. We saw before that a low bias on the reactance tube will make it act like a relatively small inductance. Decreasing the inductance in the tuned circuit of the oscillator raises its frequency. This is just what is desired if the natural frequency is low.

Turning to Fig. 8-12(C), we find the opposite case to that of (B). The relative values of the diode plate signals and the rectified voltages across *R*191 and *R*192 are reversed. The upper end of *R*191 is thus made negative with respect to the lower end of *R*192, and the bias applied to the reactance tube is increased. This causes the reactance tube to increase the inductance of the tank circuit, bringing the frequency down. This, of course, is the necessary direction of compensation when the natural frequency is high.

Additional Circuit Features

We have now covered the major points to be considered in a discussion of this circuit. However, there remain some details of interest, which we shall go over briefly.

Capacitors *C*167 and *C*170, together with resistor *R*193, form a filter to reduce the high-frequency components of the control signal coming from the sync discriminator. The discriminator circuit itself will respond very rapidly to any changes that may take place in the relative phase of the sync pulses and the sine wave generated by the horizontal oscillator. Such phase changes may be due to temporary apparent displacements of the sync pulses, as are caused by bursts of static. We have already seen, however, that these apparent displacements tend to cancel themselves out, if a large enough sequence of pulses is considered. Therefore, if the high-frequency components of the control signal are removed, the effects of short-time apparent displacements of the sync pulses will thereby be removed. At the same time, if the low-frequency and d-c components are passed through, slow variations in the frequency of the horizontal oscillator will be properly compensated. Thus, this afc circuit synchronizes the horizontal oscillator to the *average* frequency of the sync pulses, but eliminates jitter in the picture on the cathode-ray tube screen due to erratic short-time changes in the *apparent* spacing of the sync pulses.

The variable resistor *R*168 is the Horizontal Hold Control. It has a very slight effect on frequency, and allows the operator of the set to adjust the horizontal oscillator so that the afc circuit can perform its functions properly. Coarse tuning of the oscillator, required in alignment, is accomplished by means of the movable core in the primary of *T*108. The movable core in the secondary is used to balance the discriminator action. The discriminator must be balanced so that it neither adds to nor subtracts from the —2v. bias when the natural frequency of the oscillator is correct.

The cathode resistor, *R*194, of the reactance tube has an unusually low value because it is in the tank circuit of the oscillator, and a large resistance would impair the efficiency of that circuit. In addition, a voltage shifted 90° relative to the plate signal of the tube is required. As we have already seen, *R*194 must be small in order to approach this value as closely as possible. By injecting the phase-shifted signal into the cathode of the reactance tube instead of into the grid along with the control voltage, separation between these two signals is accomplished. This separation prevents undesired interaction between the oscillator and control circuits. Because *R*194 is so small, the d-c voltage across it is negligible. Hence the necessity for the —2v. fixed bias applied through the discriminator.

CHAPTER 9

SWEEP CIRCUITS

By Richard F. Koch

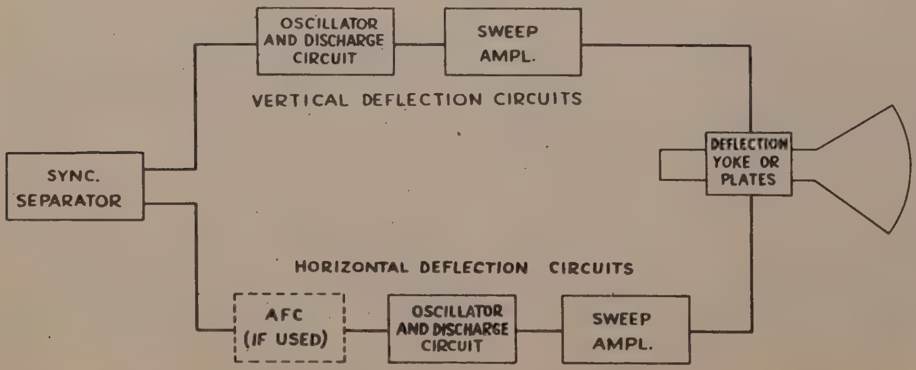
In chapter 1 we explained briefly how the electron beam in a picture tube can be deflected by saw-tooth waves applied to the deflecting elements of the tube. (More will be said on this subject in chapter 10.) The necessity for synchronizing these waves was also explained in chapter 1 and in chapter 8 we showed how the sync signals are selected at the receiver. The next problem is to generate the saw-tooth waves, control them by the sync pulses, and apply them to the sweep (or deflection) system of the picture tube. We have already seen, in chapter 8, one way in which the sync pulses may be used to synchronize the horizontal sweep oscillator.

The block diagram of Fig. 9-1 shows the essential parts of the sweep circuits used in television receivers. From the sync separator described in chapter 8, horizontal sync pulses are applied to the horizontal sweep circuit (through the afc circuit, if one is used) and vertical sync pulses to the vertical sweep circuit. Both sets of sweep circuits are basically similar; the differences which exist are due to the difference in operating frequencies. (You will recall that the horizontal sweep frequency is 15,750 cycles, and the vertical sweep frequency is 60 cycles.) The most pronounced difference appears in the deflection coils used in magnetic sweeps, and in the circuits immediately associated with them.

The saw-tooth waveforms shown in Fig. 9-2 were discussed in chapter 1. There it was pointed out that the portion of the saw-tooth wave marked "trace" is that which carries the spot from the left of the screen to the right, or from the top to the bottom, as the case may be. During the portion marked "retrace," the spot returns to the left or top of the screen. While the trace voltage or current is being applied to the deflecting system, the screen is illuminated, but during retrace it is blanked out (see chapter 1). In other words, the picture is actually being created only while the spot is moving in the trace direction. Therefore it is very important that the motion of the spot during trace be the same as the motion of the scanning spot in the camera tube. Otherwise, the relative positions of the various elements in the received picture would not be the same as in the original, and thus the received picture would be distorted. However, since the spot does not illuminate the screen during the retrace period, its exact motion at this time is not important; it is only necessary that the beginning and end of the retrace be the same as that of the spot in the camera tube. This property of the retrace is very important in magnetic scanning systems, as we shall see later.

Electrostatic vs. Magnetic Sweep

We have seen in chapter 1 that an electron beam may be deflected either by a suitable electric field set up between a pair of plates within the tube, or by the required magnetic field which is produced by a pair of coils (actually a single coil divided into balanced halves) around the neck of the tube. Both of these methods have their advantages, and both



OSCILLATOR AND DISCHARGE CIRCUIT
SWEEP AMPL.
VERTICAL DEFLECTION CIRCUITS
SYNC. SEPARATOR
DEFLECTION YOKE OR PLATES
HORIZONTAL DEFLECTION CIRCUITS
AFC (IF USED)
OSCILLATOR AND DISCHARGE CIRCUIT
SWEEP AMPL.

Fig. 9-1.—Block diagram which shows the essential portions of the sweep circuits which are used in television receivers. Horizontal and vertical sync pulses go from the sync separator to the respective deflection circuits as shown. Both sweep circuits are basically similar, the differences being due to the operating frequencies.

their disadvantages. The great advantage of the electrostatic sweep is its simplicity, which makes for compactness. (We shall see later in this chapter why the electrostatic sweep is the simpler of the two types.) Next in line is the low power requirements, particularly when it is used for the smaller picture tubes. A further advantage is that the "ion spot" difficulties associated with magnetically deflected tubes are not found when electrostatic deflection is used.

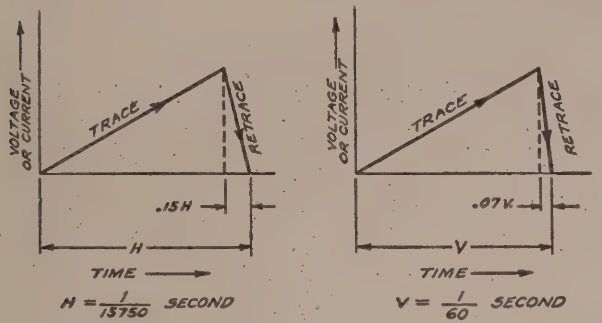


FIG. 9-2.—At the left is shown the saw-tooth waveform needed for horizontal scanning and on the right, the waveform for vertical scanning. While the "trace" voltage is applied to the deflecting system, the screen is illuminated, but during "retrace," it is blanked out.

An ion spot is a mark on the face of the picture tube produced by localized bombardment of the marked spot by negative ions. In the case of magnetic deflection, the ions are negligibly affected by the deflecting fields, but are focused in the electron gun (when electrostatic focusing is used, as is generally the case) in a manner similar to that producing the electron beam. Therefore, they proceed straight down the tube, striking the screen in the center. After prolonged bombardment by these ions, the screen is damaged, producing a defect in the picture at that point. This is called an ion spot. In an electrostatically deflected tube, the ions are deflected with the electrons, and so their effect on the screen material is spread out over the entire picture. In a magnetic-deflection system an ion trap may be used to prevent the formation of an ion spot (how this is accomplished is described in chapter 10); but of course the addition of another device to a television receiver further complicates it.

So far we have considered the advantages of electrostatic deflection. They are of sufficient weight so that this method is used almost exclusively for the smaller direct-viewing picture tubes. Compared to these advantages, the following are the advantages of magnetic sweeps. Distortion of the picture shape can be kept lower with magnetic sweeps; that is, with a magnetic sweep, the sides of the picture will be straighter than with electrostatic deflection. This effect is not so noticeable at the comparatively low anode voltages used in the smaller direct-viewing tubes; but in the large tubes and the projection tubes, both of which use high voltages, it is important. Another advantage that makes magnetic sweep desirable when high-voltage picture tubes are used is the fact that the sensitivity of a picture tube to magnetic deflection does not decrease nearly so rapidly with increasing voltages as when electrostatic sweep is employed. For this reason, the sweep voltages required for electrostatic deflection of a high-voltage picture tube are impractically high, while the magnetic fields needed for sweeping such a tube are within reasonable limits. Stating this a bit differently, we may say that in the larger picture tubes and in projection tubes very high voltages are used to accelerate the electrons in the beam, much higher than in the small cathode-ray tubes. Because of this, the large tubes require higher deflection voltages or fields, depending upon whether electrostatic or magnetic sweep is used. However, the increase in deflection voltage required by higher anode voltages is so great as to make electrostatic sweep in the high-voltage tubes impractical, while the necessary increase in magnetic field is not so great.

Sweep Oscillators

There are two main types of oscillators in use today to produce the saw-tooth waves required in deflection circuits. Both of these have nonsinusoidal outputs; that is, they do not produce sine waves. The Potter sweep circuit (named for its inventor) produces saw-tooth waves directly. The blocking oscillator either can produce saw-tooth waves directly or can be used in conjunction with another circuit, employing what is called a discharge tube, to perform this function.

(Other circuits are used, of course, besides these two types. For example, you will recall that in chapter 8 we considered a horizontal sweep afc circuit, in which the oscillator was of the Hartley type, and produced sine waves. These waves were then passed through circuits which modified them in such a way as to produce the desired saw-tooth voltages.)

The Potter sweep circuit is a special type of multivibrator. The multivibrator is an old circuit (considering the recent birth of the electronic art), consisting basically of two amplifiers, interconnected so that the output of one feeds the input of the second, and the output of the second feeds the input of the first. This is shown in block diagram form in Fig. 9-3. No frequency-selective (tuned, or resonant) circuits are included; if they were, this arrangement would give us a sine-wave oscillator. In the absence of tuned circuits, the two amplifiers do not produce a smoothly changing output, such as a sine wave. Instead, the output exhibits abrupt changes, as though it were produced by the action of a two-position switch. Actually, a multivibrator is just that—an electronic switch.

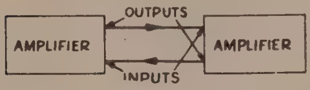


FIG. 9-3.—Block diagram of a multivibrator.

Figs. 9-4 and 9-5 show the Potter sweep circuit in its simplest form. In Fig. 9-4 this circuit has been drawn so that an easy comparison with Fig. 9-3 can be made, and in Fig. 9-5 is a diagram suited to a discussion of the operation of the circuit. Referring to Fig. 9-4, we see on the left an amplifier of the grounded-grid type, ordinarily found in v-h-f or u-h-f amplifiers only. We saw in chapter 8, under the discussion of the reactance tube used in horizontal afc, that such an arrangement is fundamentally no different from the conventional grid-input type. If the potential of the grid is fixed (by grounding it), and the potential of the cathode is varied in accordance with some signal, the grid-to-cathode voltage is thereby likewise varied. Ordinarily when a signal is applied to the grid of a tube and this signal goes negative, the flow of plate current is reduced; and when it goes positive, the plate current is increased. In this circuit, however, when the signal is applied to the cathode, a positive signal makes the grid effectively go negative with respect to the cathode, and so a positive signal decreases the plate current. When a negative signal is applied, this has the effect of making the grid-to-cathode voltage change in the positive direction; in this case, then, the plate current increases.

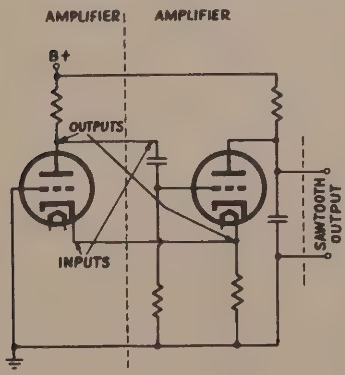


FIG. 9-4.—Simple form of the Potter sweep circuit, which is arranged to correspond with the block diagram of Fig. 9-3.

The output of the left-hand amplifier in Fig. 9-4 is taken from a resistive plate load in the usual fashion and applied to the grid of the amplifier on the right. This amplifier functions as a cathode-follower, insofar as its output which is fed to the left-hand amplifier is concerned. We thus find in Fig. 9-4 an arrangement such as that shown in the block diagram, Fig. 9-3.

Let us now turn to Fig. 9-5 and consider the manner in which this circuit operates. To begin with, let us assume that the heaters have been turned on, but not the plate supply. This being the case, neither C_1 nor C_2 is charged. When the plate voltage is turned on, several actions occur at the same time. These actions are all dependent upon each other, but the one that may be considered basic is the charging of C_1. This charges through R_1, the cathode-grid circuit of V_2, and R_4. (An appreciable charging current can flow from the cathode of V_2 to its grid because the grid is made positive at this time with respect to the cathode. The grid, therefore, acts like a diode plate for the cathode, permitting the flow of current necessary to charge C_1) Some charging current will also flow through R_3, but this contribution is negligible, since the resistance of R_3 is much higher than the resistance of R_4 and the grid-cathode resistance of V_2 (when the grid is positive). The charging current through R_1 lowers the plate voltage of V_1 below the supply voltage, and the flow of this current through R_4 raises the cathode voltage of this tube to a value considerably above ground. The combination of these two factors—low plate voltage and high cathode bias—causes V_1 to be cut off. This condition obtains immediately after the plate voltage is turned on, but it lasts for only a very short time, as we shall see.

As soon as C_1 starts to charge, that is, as a voltage builds up across it, the plate voltage of V_1 rises. At the same time, the charging current falls off. (You will recall that when a capacitor is charged through a resistor, maximum current flows when the circuit is first completed. Thereafter, as the voltage across the capacitor increases, the charging current decreases.) Since this charging current passes through R_4, the voltage across this resistor decreases with the decreasing current through it. Thus, after the first instant, in which V_1 is cut off by the low plate voltage and high cathode bias voltage existing on it, the plate voltage rises and the cathode voltage falls.

It takes only a very short time for these potentials on tube V_1 to reach such values that V_1 will begin to conduct plate current. When this happens the voltage drop that occurs across R_1 is transferred through C_1 to the grid of V_2. Let us consider just how this transfer occurs. When V_1 begins to conduct, this has the effect of suddenly applying a negative voltage (in series with whatever voltage already existed) to the upper plate of C_1. When this happens, C_1 must begin to discharge, since the charge which it already had was such that its upper plate was positive. In order to discharge C_1 electrons must flow away from its lower plate. They cannot flow from the grid of V_2 to the cathode (in any appreciable quantity), because of the diode action that occurs there. However, they can flow through R_3. This flow makes the end of R_3 to which C_1 and the grid of V_2 are attached negative with respect to ground. Thus we may say that the sudden voltage appearing at the plate of V_1 is transferred through C_1 to the grid of V_2.

While C_1 is charging, the grid of V_2 is positive with respect to its cathode, as we mentioned before. Therefore, at this time V_2 is conducting. This puts a partial short circuit across C_2, and prevents it from acquiring any appreciable charge. However, when the grid of V_2 is driven negative by the conduction of V_1, the plate current in V_2 is greatly decreased. The voltage drop across R_4 due to the plate current of V_2 likewise decreases. Since the cathode of V_1 is connected to the upper end of R_4, this condition applies a negative signal to this cathode. (Since the upper end of R_4 is less positive, it is going *negative.*) As we saw previously, a negative signal at the cathode of a tube causes the plate current to increase. Therefore, the drop across R_1 increases, and this makes the upper plate of C_1 go still further in the negative direction. This negative signal is transferred to the grid of V_2 as before. The cathode of V_2 then becomes still less positive. This negative signal from the cathode of V_2 causes a further increase in the plate current of V_1. Thus we have a cumulative action; the decrease in the plate voltage of V_1 puts a negative signal on the grid of V_2, which transfers a negative signal to the cathode of V_1, causing an increase in plate current in V_1 and a drop in plate voltage, which puts a further negative signal on the grid of V_2, etc., etc. This cumulative action continues until V_2 is cut off; after this the plate current of V_2, being zero, cannot decrease further; and the plate signal of V_1 can no longer be transferred back to its cathode. When V_2 is cut off, the near short circuit that existed across C_2 when V_2 was conducting is changed to an open circuit. This being the case, C_2 charges through R_2. The switching of V_2 from the conductive condition takes place almost instantaneously.

It was pointed out previously that the transfer of plate signal from V_1 to the grid of V_2 through C_1 is due to the discharge of C_1 through R_3. This discharge takes an appreciable length of time; this time is determined by the "time constant" of the circuit. The time constant of a series RC circuit is a measure of the length of time required for the capacitor to lose a certain percentage of its charge; in such a circuit it is numerically equal to the product of the resistance in ohms and the capacitance in farads. In this case, the resistance in series with the capacitance is R_3, R_4, and the plate resistance of V_1. In practical design, R_3 is usually much the largest of these three resistances, so the discharge time of C_1 is dependent upon the product R_3C_1. While this discharge takes place, a negative voltage exists on the grid of V_2. In this connection, another fact should be remembered, namely, that as the discharge proceeds, the amplitude of the discharge current decreases. As this current decreases, the voltage across R_3 due to it also decreases. Since this voltage is the bias on V_2, a point in this decrease will be reached where V_2 is no longer cut off. At this point, of course, V_2 will begin to conduct. Another factor in determining when V_2 again becomes conductive is the charging of C_2 through R_2. As C_2 charges, the plate of V_2, which is connected to it, becomes more and more positive. Because of this, it conducts sooner than if the plate voltage were constant, and only the grid voltage changed.

When V_2 starts to conduct, C_2 discharges through

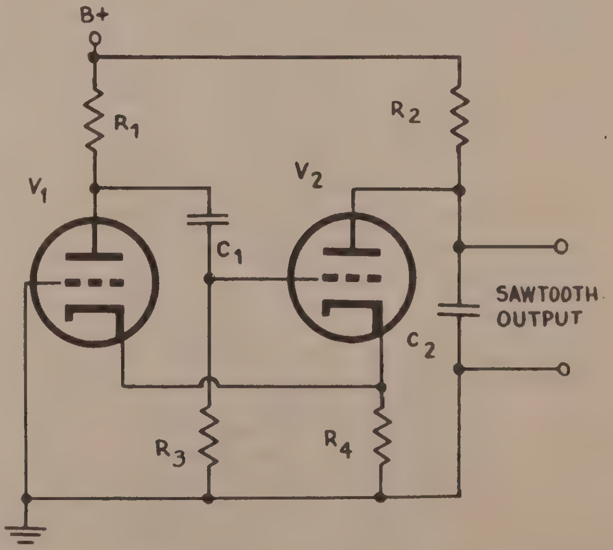


FIG. 9-5.—The Potter sweep circuit. The output of the left-hand amplifier is fed to the grid of the second amplifier V_2, which functions as a cathode follower, as far as its output that is fed to V_1 is concerned.

it, causing a much larger current to flow through R_4 than when V_2 is cut off. The resultant large voltage drop applies a positive signal to the cathode of V_1. As we have seen, such a signal causes a decrease in plate current. A decrease in the plate current of V_1 will reduce the drop across R_1. This sudden application of a positive potential to the upper plate of C_1 causes it to charge through the grid-cathode circuit of V_2 as it did when the plate supply voltage was first turned on. In order for this charge path to exist, the grid of V_2 must, of course, be positive with respect to the cathode. Under this circumstance, the tube conducts very heavily in its plate circuit, discharging C_2 rapidly. When C_2 is discharged, the current through R_4 drops to a relatively small value. (R_2 is a relatively large resistor, so only a small current can flow through it.) The resultant drop in voltage across R_4 applies a negative signal to the cathode of V_1. This makes the plate of V_1 go negative. We have thus arrived at the same condition as existed a very short time after the plate supply voltage was first turned on. That is, C_1 has been charged and V_2 is conducting. As before, the charging current C_1 decreases as its charge increases, causing a smaller drop across R_4 than existed when C_1 first began to charge. The sequence continues with V_1 becoming conductive and V_2 being cut off, allowing C_2 to charge up once more. Thus C_2 is charged slowly through R_2 and discharged quickly through V_2 and R_4, giving the saw-tooth waveform shown in Fig. 9-2.

When this circuit is used in a television receiver or an oscilloscope, the grid of V_1 is grounded through a resistor, instead of directly. Sync signals can then be injected into the circuit across this resistor. Another minor change sometimes found is the use of a dual triode tube having a common cathode (such as the 6J6), instead of two separate triodes.

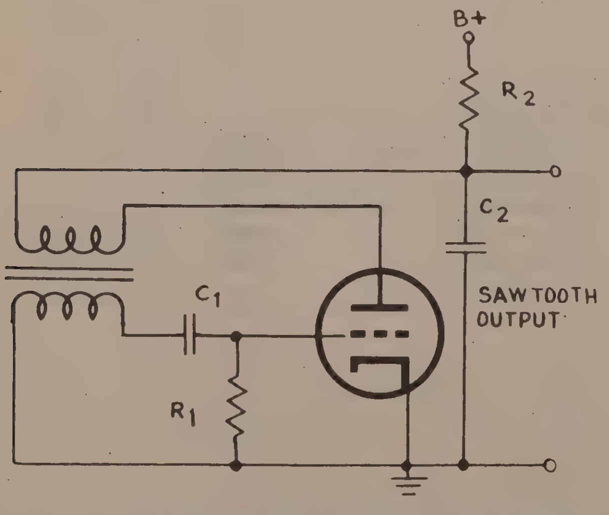


FIG. 9-6.—A saw-tooth wave is obtained from the plate circuit of this blocking oscillator and very sharp pulses may be derived from the grid circuit.

The Blocking Oscillator

In Fig. 9-6 is shown a very simple form of the blocking oscillator. In addition to the saw-tooth wave obtainable in its plate circuit, very sharp pulses may be derived from the grid circuit. This property is of great importance, and considerably broadens the utility of this oscillator. For example, it is widely used as a pulse generator in radar equipment. In addition, as we shall see later, these pulses can be very important in sweep circuits.

In order to see how the blocking oscillator operates a very important observation should be made, namely, that this circuit has the same form as the common tickler feedback oscillator, with the feedback transformer tuned by its own distributed capacitance. The difference between the two oscillators lies in the choice of values for R_1 and C_1. In the usual tickler feedback oscillator, as used, for example, in a superheterodyne receiver, the oscillator operates class C, and C_1 and R_1 are used to generate grid bias. The oscillator generates a continuous signal at the resonant frequency of the feedback circuit; at the positive peak of each cycle a little grid current is drawn, putting a charge on C_1. This charge leaks off between positive peaks through R_1, biasing the grid negative. The strength of this bias is dependent upon several factors, among them the values of C_1 and R_1.

In the blocking oscillator, C_1 and R_1 are deliberately chosen so large as to interfere with the continuous production of sine-wave oscillations at the resonant frequency of the feedback transformer. Instead of a continuous train of such oscillations, a series of pulses, each corresponding to a positive peak of the suppressed sine waves, is formed. The frequency of these pulses is determined primarily by R_1 and C_1. How this is accomplished, and how a saw-tooth wave is formed in the plate circuit, we shall now see.

Let us begin, as in the case of the Potter sweep circuit, by assuming that the heater of the tube is lit, but the plate supply is not yet turned on. This being the case, C_1 and C_2 will not be charged. When the plate supply is turned on, plate current will start to flow in the tube. (C_2 will start to charge at this time, but this effect is very slight, because the resistance of the plate winding of the transformer in series with the plate resistance of the now heavily conducting tube is much smaller than R_2.) Since plate current is now flowing, where an instant previously it was not, there has been a *change* in the current through the plate winding of the transformer. This change in current is coupled into the grid winding, appearing there as a voltage across the winding. The polarity of this voltage is such as to make the grid positive with respect to ground (and the cathode). This causes the tube to conduct still more heavily, causing a further increase or change in the same direction in the plate winding, causing the grid to become more positive, causing the plate current to increase still more, etc., etc. At the same time grid current is being drawn, piling up electrons on the plate of C_1 nearer the grid, thereby charging this capacitor. The cumulative action which we have described—the interdependent increase of plate current and grid voltage—takes place very rapidly, at a rate determined primarily by the resonant frequency of the transformer. However, it cannot go on indefinitely. As the plate current increases, it reaches a value where the rate of increase drops off; in fact, it tends to stop increasing altogether. This tendency is due to limitations imposed by circuit impedances and tube characteristics.

When this happens, it produces a major reaction in the grid winding of the transformer. You will recall that the voltage across the grid winding is determined by the *change* of current in the plate winding. Up to now, the current has been increasing, but now it is starting to decrease. Therefore, the direction of *change* has *reversed* (although the *current itself* is still flowing in the *same* direction). This causes the voltage across the grid winding to reverse also. The grid is now suddenly driven negative, and the flow of plate current is decreased still further by this action. Now another cumulative action is set up; this time the decrease in plate current drives the grid negative, which further decreases the plate current, which drives the grid further negative, etc., etc. In this manner the tube changes very rapidly from a condition of heavy plate-current conduction to one of cutoff.

When the grid is driven negative, no appreciable number of electrons can flow from it to the cathode. However, a large number of electrons, which flowed through the low-resistance circuit that existed from cathode to grid when the latter was positive, are piled up on the grid side of the capacitor C_1. With the reversal of voltage across the grid winding of the transformer, these electrons are driven away from that plate by the negative potential placed at the opposite plate. The only path open to them is through the high resistance R_1. While electrons flow from C_1 through R_1 to ground (and back to the transformer side of C_1) to discharge the capacitor, the grid is kept negative. Since R_1 is a much higher resistance than the cathode-to-grid resistance of the tube *when* the grid is *positive*, a much longer time is required to discharge C_1 than is required to charge it. This holds the grid negative for a much longer time than it is positive.

The negative voltage applied to the grid cuts off the tube, removing the near short circuit that was placed across C_2 when the tube was conducting. C_2 is thus allowed to charge through R_2, and so a slow, steady rise of voltage takes place across it. This of course makes the plate voltage more and more positive; no plate current is flowing, so there is no voltage drop across the plate winding of the transformer, and therefore the plate voltage is equal to the voltage across C_2. At the same time, the discharge current from capacitor C_1 is steadily dropping, since the charge on the capacitor is falling off. In turn, the voltage drop across R_1 due to the discharge current also decreases, and thus the grid bias becomes less and less. Finally, the grid bias becomes low enough, and the plate voltage high enough, so that plate current can flow.

As soon as plate current begins flowing, a positive voltage is applied to the grid through the transformer, increasing the flow of plate current—and thus the chain of events that occurred after the plate voltage was first applied is repeated. Once more a near short circuit is placed across C_2, discharging it through the plate circuit of the tube. At the same time, C_1 receives a large charge through the grid circuit. As before, this takes place very quickly, and then the tube returns to cutoff, in which condition it remains a comparatively long time. Thus, C_2 is allowed to charge for a long time, but slowly; when it is discharged, this operation takes but a fraction of the charge time, consequently the rate of discharge is much greater than the charge rate. This produces the desired saw-tooth wave across C_2.

In actual practice there are two modifications to the circuit shown in Fig. 9-6. In this figure one end of the grid winding of the transformer is shown grounded; ordinarily, however, it is connected to a source of sync signals. The other modification, although frequently used, is not nearly so universal as the first. This is the use of an additional triode, connected as in Fig. 9-7 (a sync input is also shown here). R_4 in Fig. 9-7 completes the d-c path in the grid circuit of the blocking oscillator, this path having been broken to permit injection of the sync signals. C_4 shunts R_4, so that the a-c impedance of the grid circuit is not considerably increased by the addition of the sync circuit. Without C_4, R_4 would increase the charging time of C_1. Since the charging time of C_1 corresponds to the retrace part of the saw-tooth wave shown in Fig. 9-2, this would have an undersirable effect, for it is necessary that the retrace be very short, as shown.

Looking at Fig. 9-6, we see that the discharge path of C_2 lies in part through the plate coil of the transformer. This inductance has a very pronounced retarding effect upon the rate of discharge. Therefore, if the discharge is to take place quickly, only a partial discharge of C_2 can be effected in the short time allowed. The result of this will be a saw-tooth wave with a very low amplitude. Conversely, if a large amplitude is required, that is, if the charge and discharge of C_2 must cover a large range of voltage, a relatively long time must be allowed for the retrace.

To get around this difficulty, a second tube may be used as a switch to charge and discharge the capacitor across which the saw-tooth waveform is produced. Such an arrangement is shown in Fig. 9-7. Here the left triode is connected in the same manner as the triode in Fig 9-6 (except for the addition of the sync input circuit). The grid of the right triode is connected to the grid of the left triode, and so they are cut off and conductive simultaneously. Since the saw-tooth output is no longer obtained across C_2, the saw-tooth amplitude here is not important. Therefore the retrace time can be made short. On the other hand, since there is no inductance in the discharge path of C_3, a large part of the charge on C_3 can be removed very quickly. Thus a saw-tooth wave having a high amplitude and rapid retrace is produced by the use of the extra triode.

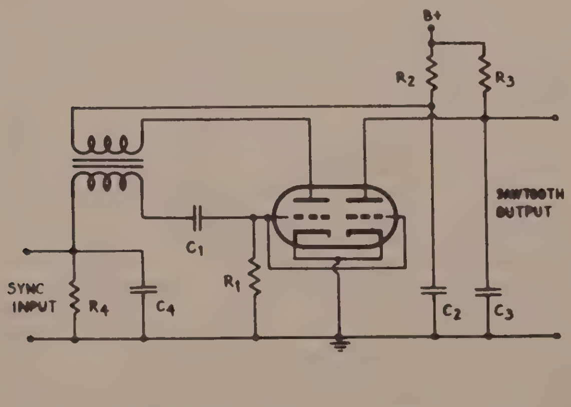


FIG. 9-7.—The addition of the second triode in this blocking oscillator enables the production of a saw-tooth wave having a large amplitude and rapid retrace.

Required Sweep Waveforms

Fig. 9-2 shows the manner in which the position of the scanned spot on the picture tube must change with time. If electrostatic deflection is used, the final sweep amplifier can be connected to the deflection plates as shown in Fig. 9-8. The loads on the final sweep amplifier tubes (their plates only are shown, at the left) are purely resistive; the effects of the d-c blocking capacitors and the input capacitance of the picture tube are negligible. Since the deflection of the spot in the picture tube is directly dependent upon the voltages applied to the deflection plates, a saw-tooth voltage of the shape shown in Fig. 9-2 is required to obtain this same saw-tooth variation of position. Therefore, since the amplifier loads are resistive, their plate currents and grid voltages must be of the same shape as their plate voltages, namely, a simple saw-tooth wave. This permits the use of a simple sweep generator such as the Potter sweep circuit or the blocking oscillator, and the waveform needs no modification after it is generated. The over-all sweep system (sweep oscillator plus amplifiers) for use with an electrostatic-deflection picture tube is, therefore, rather simple. Let us now compare this system with those required for magnetic deflection. We shall see that the vertical deflection waveforms are more complicated than those used in an electrostatic-deflection system, and the horizontal even more so.

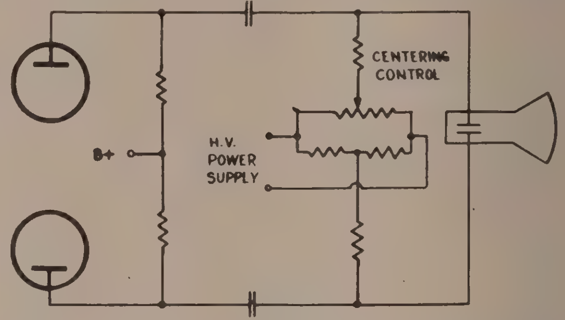


FIG. 9-8.—When electrostatic deflection is used on the picture tube, the final sweep amplifier can be connected to the deflection plates as shown in this schematic.

Typical Sweep Circuit for Magnetic Deflection

Fig. 9-9 shows the vertical deflection circuit of the Stromberg-Carlson TV10P Series 10 and 11 television receivers. In order to simplify the illustration, connections to different points in the positive low-voltage supply are distinguished by marking one with a prime and the other with a double prime. The 6N7 tube operates in a blocking oscillator circuit of the type shown in Fig. 9-7. Although there are several differences between the circuits, there is only one that is more than a matter of detail. Let us consider these differences, disposing of the simple ones first.

In the first place, there are the variable resistors. In discussing Fig. 9-6 we saw that the sweep frequency is determined principally by C_1 and R_1, and this is also true of the circuit of Fig. 9-7. In Fig. 9-9, the corresponding resistance (*R*-136 and *R*-137 in series) is variable. This permits the operator of the receiver to adjust the natural frequency of the oscillator so that the sync pulses can *hold* it at exactly the right frequency. Hence the name VERT. HOLD. Potentiometer *R*-140 controls the voltage available to charge the capacitor *C*-128 across which the sweep saw-tooth wave is formed. It therefore functions as the VERT. AMPLITUDE control.

No d-c path to ground from the transformer side of *C*-127 is shown. However, this path does exist, through the plate circuit and the plate voltage supply of the second sync amplifier.

We now come to the significant difference, the resistor (*R*-144) in series with the capacitor (*C*-128) across which the sweep signal is generated. Although the lower plate of this capacitor is not directly grounded, as in the circuits of Figs. 9-6 and 9-7, it is almost at a fixed potential (the equivalent of ground for the sweep signal frequency) because of the large bypass capacitor *C*-126-*D*. Actually, a very small signal exists here. This serves as a type of negative feedback to improve the characteristics of the sweep signal. (We say "a type of" because the nonsinusoidal input to the 6SN7 is distorted by capacitor *C*-126-*D*. However, the nature of this distortion is such as to give a signal which helps to cancel some of the distortion which exists in the input to the 6SN7 in the first place.)

To determine the purpose of *R*-144 we must first consider the properties of the vertical sweep coil. The coil has both inductance and resistance. This is a very important point. Another important point is that in a magnetic sweep system, saw-tooth deflection of the electron beam is accomplished by passing a saw-tooth *current* through the deflection coil. This is different from an electrostatic-deflection system, in which a saw-tooth *voltage* is required across the deflection plates to give a saw-tooth sweep. Now let us look at Fig. 9-10. At (A) is shown the saw-tooth current wave required to give saw-tooth deflection on the beam. At (B) we see an inductance and resistance in series, representing the actual inductance and resistance that exist in a deflection coil. In an actual coil, of course, the two components are not separated as in Fig. 9-10(B), but this representation is convenient for our discussion and does not introduce any errors. Using this convenient separation of the resistive and inductive components of the coil, we can determine the voltage waveform across the coil when a saw-tooth wave of current is passed through it.

In Fig. 9-10 (C) are shown three waveforms. As implied by their captions, they are related to

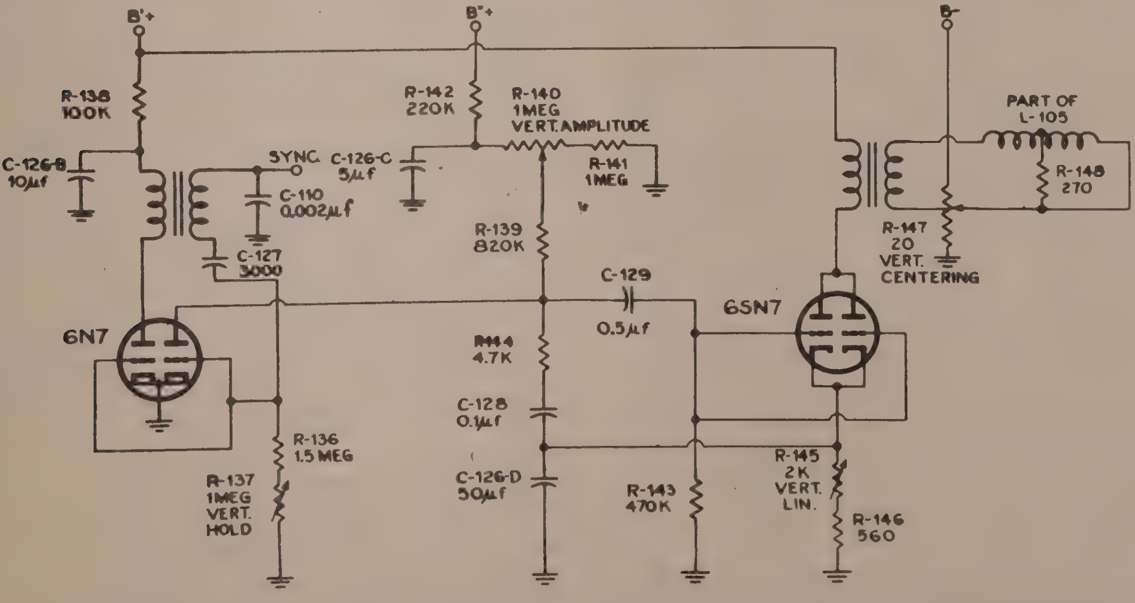


Fig. 9-9.—The vertical deflection circuit of the Stromberg-Carlson model TV10P, Series 10 and 11.

the waveform at (A) and the coil at (B). We know from both theory and experimental observation that when a steadily changing current (that is, a current having a linear rise or fall) is passed through an inductance, it produces a fixed steady voltage across the inductance. This voltage is proportional to the rate at which the current is changing and to the size of the inductance. Since our inductance here is fixed in value, any change in the voltage waveform across it must be due to a change in the current through it. And the current does change; first it rises fairly slowly, that is, it changes in the positive direction. As a result, there is a positive voltage across the inductance which endures for exactly the same length of time as the current rises. Then, abruptly, the current starts to decrease, that is, it changes in the negative direction. At the time that this change in the current wave takes place, there is a corresponding change in the voltage wave; and since the current is changing steadily in the negative direction, the voltage is steady at a negative value. The duration of the negative swing of the voltage is exactly the same as the duration of the negative change of current that produces it. Thus alternating fixed positive and negative voltages are produced by the passage of a saw-tooth wave of current through an inductance. This voltage waveform is sometimes known as a rectangular wave; it is also sometimes called a "square wave," but it is more exact to restrict this latter term to a wave similar to that shown in which the durations of the positive and negative swings are *equal*. This usage stems from the distinction between a rectangle and a square; in both all the angles in the corners of the figures are right angles, but in the square there is the additional specification that all sides are *equal*.

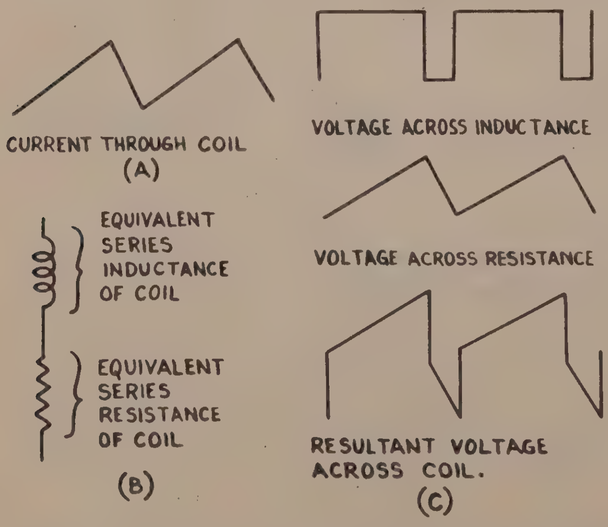

CURRENT THROUGH COIL
(A)
VOLTAGE ACROSS INDUCTANCE
EQUIVALENT SERIES INDUCTANCE OF COIL
VOLTAGE ACROSS RESISTANCE
EQUIVALENT SERIES RESISTANCE OF COIL
(B)
RESULTANT VOLTAGE ACROSS COIL.
(C)

Fig. 9-10.—When a saw-tooth current (A) flows through a coil, the combined waveforms of the voltages due to the inductive and resistive properties of the coil (B) produce the resultant waveform shown at the bottom of (C).

The voltage waveform across the resistive component is the same as the current waveform that produces it. This follows from Ohm's law: $E = IR$. Since the resistance is fixed, the voltage must vary in exactly the same manner as the current.

When a rectangular wave and a saw-tooth wave are combined as shown in Fig. 9-10 (C), the resultant is known as a trapezoidal wave. This, then is the voltage wave that must exist across the deflection coil when the necessary saw-tooth current passes through it. In the coils ordinarily used, the resistance is small, and so the resistive component of voltage is much smaller than the inductive. The resultant trapezoidal wave, therefore, does not have as pronounced slopes as does that shown in Fig. 9-10(C), but is almost a rectangular wave. However, its difference from a simple wave is sufficiently great to be significant.

Having determined the voltage waveform that must exist across the coil in order to achieve the desired saw-tooth deflection, our next problem is how it is generated. This is where the mysterious resistor, *R*-144 (see Fig. 9-9), enters the picture. Fig. 9-11 illustrates its function. Part (A) of this figure indicates the charging current of *C*-128 as a positive current, which flows for a relatively long time, and the discharge current as a negative flow having a short duration. The combination of *R*-144 and *C*-128 is shown at (B) for ready reference. In part (C) are found the individual voltages existing across the components shown in (B) and the resultant voltage.

As we have already pointed out, it follows from Ohm's law that the waveform of the voltage across a resistor is the same as the waveform of the current through it. This accounts for the rectangular voltage wave across *R*-144. We have also seen that a steady flow of current charging or discharging a capacitor produces a steadily rising or falling voltage across it. Hence the saw-tooth wave of voltage across *C*-128. The resultant of these two waves is a trapezoidal wave, just as in Fig. 9-10(C). Thus the addition of *R*-144 modifies the output of the blocking oscillator in the manner required for a magnetic sweep.

Another interesting feature of the circuit shown in Fig. 9-9 is the VERT. LIN. (vertical linearity) control. The function of this adjustment is to make the vertical deflection of the electron beam as nearly perfect a saw-tooth wave as possible. The waveforms shown in Figs. 9-10 and 9-11 are, of course, idealized. Actually, there will be some curvature in them; but if distortion of the picture on the face of the cathode-ray tube is to be minimized, this curvature must be made as slight as possible.

To reduce the curvature, advantage is taken of a characteristic of vacuum-tube amplifiers that is usually a source of trouble. This characteristic is the curvature of the grid voltage-plate current characteristic, which causes amplitude (or harmonic) distortion. By good fortune, this curvature is opposite to the curvature of the sweep waveform generated by the blocking oscillator, and so the two tend to cancel. You will recall that the amplitude distortion produced in an amplifier is dependent on the bias, and that the bias can be controlled by varying the cathode resistance. *R*-145 (the VERT. LIN. control) accomplishes this function, and so permits adjustment of the curvature of the 6SN7 characteristic to provide the best compensation for the curvature of the output of the blocking oscillator.

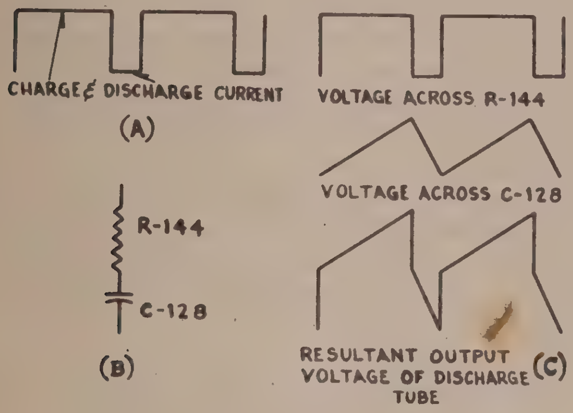

CHARGE & DISCHARGE CURRENT
(A)
VOLTAGE ACROSS R-144
VOLTAGE ACROSS C-128
R-144
C-128
(B)
RESULTANT OUTPUT VOLTAGE OF DISCHARGE TUBE
(C)

FIG. 9-11.—When a rectangular current wave (A) is applied to a resistor and capacitor in series (B), the individual waveforms are as shown in (C) with the resultant waveform at the bottom.

The output of the 6SN7 is coupled to the vertical deflection coil (part of the deflection yoke, *L*-105) through a transformer. This transformer provides an impedance match, just as an output transformer in an audio amplifier matches the voice coil of the speaker to the plate circuit of the final stage. In the secondary circuit of the output transformer we find the VERT. CENTERING control. This is a small potentiometer (20 ohms) with a fixed center tap as well as the movable arm. Current is drawn through it by one of the power supplies, and by proper adjustment of the control, part of this current can be made to flow through the deflection coil. Which way this current will flow depends upon whether the movable arm is above or below the fixed tap, and how much flows depends upon how far the arm is from the tap. In this way a fixed component of current through the coil can be added to the sweep current, and this determines where the center of the picture will lie in the vertical axis. Hence the name of the control.

Resistor *R*-148 is a damping resistor, used to reduce the tendency of the vertical deflection coil to produce unwanted oscillations. When a sharply changing voltage wave (such as the trapezoidal wave used here) is applied to coil, there is a likelihood that oscillations will be produced, since the coil is tuned by its distributed capacitance. This effect is known as ringing.

The GE model 802 television receiver employs magnetic deflection of the electron beam in the picture tube. The vertical deflection circuit is quite dissimilar in detail to that used in the Stromberg-Carlson sets just described. However, the principles are very much the same. For example, the output tube is a 6V6-GT in the GE set, instead of a 6SN7 with its two sections in parallel as in the Stromberg-Carlson receivers. But in both circuits, the vertical linearity control is located in the cathode circuit of this stage; and in both there is a small corrective signal fed back from the cathode to the grid. A somewhat more pronounced difference is the use of a multivibrator in the GE receiver as the sweep generator; and another is the manner in which a trapezoidal sweep waveform is obtained. Here the sweep generator produces an unmodified saw-tooth wave, and this is applied without change (except for the small corrective feedback) to the grid of the output amplifier. However, in the plate circuit of this tube an effect is emphasized which is carefully avoided in the Stromberg-Carlson receivers. This is ringing in the inductive load (output transformer and deflection coil.) It is emphasized by the addition of a capacitor across the secondary of the transformer.

If you will look once again at the rectangular waves shown in Figs. 9-10 and 9-11, you will see that they are in the nature of pulses. This resemblance would be even more distinct if the duration of the negative swing were a good bit shorter. On the basis of this, it may be said that a trapezoidal wave can be produced by adding suitable pulses to a saw-tooth wave. This is done in the GE model 802. During the retrace time the grid voltage of the output tube (*V*20, see Fig. 9-12) changes from its maximum positive voltage to its maximum negative voltage. This negative voltage swing is sufficient to cut the tube off momentarily and cause ringing. The capacitor (*C*44) across the secondary of the output transformer (*T*20) is rather large (0.1 μf), so that a half cycle of the ringing frequency lasts long enough for the output tube to return to its conductive state. When the tube conducts, its plate resistance is sufficiently low to damp out further oscillation. In this manner a good approach to a trapezoidal wave is obtained, using half sine-wave pulses instead of flat-topped straight-sided pulses in combination with a saw-tooth wave. This deviation from the ideal trapezoidal shape of the sweep voltage produces some distortion of the sweep. However, this distortion occurs during the retrace time, when the beam is blanked, and so its effect on the picture is negligible.

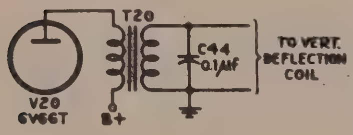


Fig. 9-12.—In the GE model 802, this circuit is used to produce a trapezoidal waveform by adding suitable pulses to a saw-tooth wave.

Ringing is put to similar use in the horizontal sweep. This is quite common practice in modern television receivers using magnetic sweep, but it has been found necessary in this case to use a separate tube to provide damping action. The necessity for the separate tube lies in the much higher sweep frequency used in the horizontal circuit, 525 times the vertical sweep frequency. The damping tube and its associated components are shown in Fig. 9-13. This circuit is treated at some length in chapter 11, but there are certain points in connection with its operation that it is well to mention here. (This circuit is discussed under power supplies because the sinusoidal pulse produced by ringing in the deflection coil is stepped up in an autotransformer to a very high voltage. It is then rectified, and provides 8200 volts for the second anode of the picture tube.)

The HORIZONTAL LINEARITY control changes the time constant of the circuit consisting of capacitor *C*94 and resistors *R*90 and *R*115. In this manner the shape of the signal applied to the grids of the damping tube is modified, and this in turn affects the shape of the sweep signal during and immediately after the retrace. Resistors *R*120 and *R*121 prevent the grid current in the damping tube from rising to an excessive value when the positive pulses due to ringing are applied to it. The WIDTH control is simply a variable inductance, similar to the deflection coil and in series with it. Because of the series connection, part of the sweep voltage appears across one coil, and part across the other. By varying the inductance of *L*23 the operator of the set can change the proportion of the sweep signal appearing across the deflection coil and, therefore, the amplitude of the horizontal sweep.

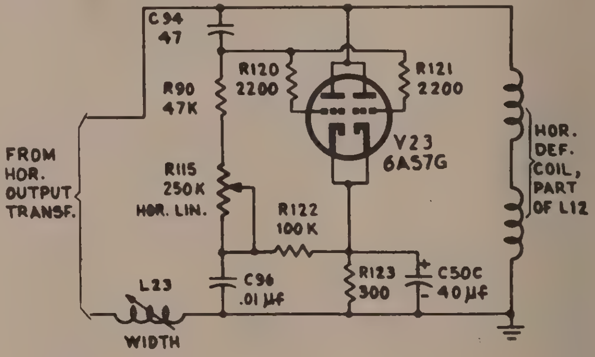


Fig. 9-13.—The damping tube and its associated components in the horizontal sweep circuit of the GE model 802.

CHAPTER 10

THE PICTURE TUBE

By William Bouie

As stated in chapter 1, the picture tube which displays the televised images has essentially the same design as the conventional cathode-ray tube. For those who desire to investigate further the theory and functioning of cathode-ray tubes, reference is made to Rider's The Cathode-Ray Tube At Work, which contains a wealth of material on these tubes. Our discussion here will be limited to the cathode-ray tube as used in television reception.

Two types of picture tube are in popular use at the present time. These are the direct-viewing-type and the projection-type tube. In the direct-viewing tube, the received image is displayed on the face of the tube and is viewed directly. In the projection-type tube, the image on the tube screen is directed through a lens system and is then projected upon a large viewing screen. Both of these tube types will be discussed in this chapter.

The type of picture tube used for direct viewing varies in size from the 5-inch tube which is incorporated in the small table-size models to the 20-inch tube used in the large console television receivers, with even larger tubes being used for experimental work. An illustration of a 10-inch picture tube, a popular size in the moderate-priced television receiver, is shown in Fig. 1-5.

Screens

The luminescent screen material which is coated on the inner surface of the large end of the picture tube performs a very important function in the operation of the tube. It is this material that converts the electrical energy in the accelerated electron beam into light whose intensity varies with the intensity of the television signal modulating the electron beam. The efficiency of this conversion is largely dependent upon the type of material used in the formation of the screen, and the thickness of the coating. Many different types of oxides and phosphors are used in forming the screens of cathode-ray tubes and each has its value in specific applications. The screen can be made to glow for long or short periods of time (persistency), usually in shades of white, green, or blue. These screens are used in test oscilloscopes and other devices used in studying and photographing observed electrical phenomena.

The screens used in television picture tubes are formed from a coating of phosphor having a white or bluish-white cast. These screens have a low persistency, the glow ceasing when the electron beam is removed from a point. This low persistency characteristic is an absolute requirement in order to prevent blurring of moving objects in the television picture. As stated in chapter 1, the screen material is usually coated with a thin aluminum backing which increases the brightness of the television image and stops the heavy ions in the electron beam, thereby preventing formation of the ion spot. The formation of the ion spot will be discussed later in this chapter.

Persistency of Vision

The electron beam causes a fluorescent spot to appear upon the screen only at the point where the beam strikes the screen. It is this *single spot,* acting under the influence of the deflecting fields and the television signal on the grid, that traces the picture on the face of the tube.

The action of the deflecting fields is such that the electron beam is made to shift its position by small increments and to produce a fluorescent spot which appears periodically at every point on the screen. This spot varies in intensity with the signal voltage applied to the grid of the tube, and rapidly traces the television picture. The progressive appearance of the fluorescent spot at all points on the screen creates the illusion of an instantaneous picture by persistence of vision and, while it is true that the fluorescent spot exists at any one point for a very small fraction of an instant, it seems to the observer that the entire area of the screen is made fluorescent at the same instant. In other words, the entire picture on the tube appears to occur instantaneously instead of being constructed of a series of infinitely small fluorescent spots of light moving with great rapidity horizontally and vertically across the face of the tube.

This persistency of vision, which is a characteristic of the human eye, is an optical illusion created by the inability of the eye to respond to changes which occur faster than a certain rate, usually one-tenth of a second. Persistence of vision can be illustrated by whirling a lighted flashlight in a circular motion in a dark room. As the speed of rotation is increased beyond a certain point, a ring of light will be observed in the darkness. The circular speed of the flashlight and the inability of the eye to spot the light beam at any one position, because of its rate of travel, causes the illusion described as persistence of vision. The motion picture is, itself, an optical illusion. Actually the subjects on the screen do not move, although the eye sees continuous motion. What really happens is that a series of still pictures is presented, each of which is stationary for a very short time, usually 1/24th of a second. The eye cannot follow change at this speed, hence the motion appears continuous.

Focusing

Fig. 10-1 shows the elements contained in the picture tube. These elements, or electrodes, all influence the stream of electrons in the formation of the television picture. Upon application of the proper voltages, electrons are emitted from the cathode and they spread outward in all directions. In order to form the spot of light which is used to trace out the image on the tube screen, these electrons must be concentrated into a beam. This action is called focusing the electron beam. Proper acceleration of these electrons is also required so that when they strike the viewing screen the impact will be sufficiently great to cause fluorescence. Focusing is also a function of this acceleration.

If the emitted electrons are not concentrated into a beam, the image position will not be under control. Neither will it be possible for the electrostatic or magnetic deflection fields, whichever are used, to act uniformly to produce an image of value unless the beam is properly focused. An incorrectly focused beam results in a large spot with a halo around it. In some instances, improper focusing results in a very wide spot which, upon being deflected, produces extremely wide lines of non-uniform width on the screen.

Optical Analogy of Focusing

Suppose that we consider a simple optical system, such as that shown in Fig. 10-2. The variables which present themselves in this analogy are very much like the variables which are found in the television picture tube. This will become evident as we continue the discussion. This optical system consists of a source of light, *L*, a stop, or light gate, *S*, two lenses, *O* and *P*, and a screen *Sc*.

The purpose of the system is to project the light from the lamp, *L*, upon the screen, *Sc*, in the form of a small but intense spot of light. The nature of the light gate, or stop, is like that of a stop on a camera; that is, it is a device with an opening of variable size. By means of this device the amount of light passed from the lamp, *L*, to the lens, *O*, is definitely under control. At the same time, we can further assume that the intensiy of lamp *L*, is also under control. In other words, a rheostat can be assumed to exist in the power supply circuit feeding the lamp.

The two lenses are of the type which will pick up the light from some source, concentrate the beam, and project a pin point of light upon a screen at a definite distance from the lens. In the process of projection from lens *P*, the light from lens *P* is caused to converge into a spot upon the the screen.

In projecting the spot, let us consider the first requirement, namely, the intensity of the spot. This, quite obviously, depends upon the amount of light available from *L*. If the original light source is not sufficiently luminous, all the concentration in the world is not going to produce a sufficiently luminous spot. On the other hand, although the light

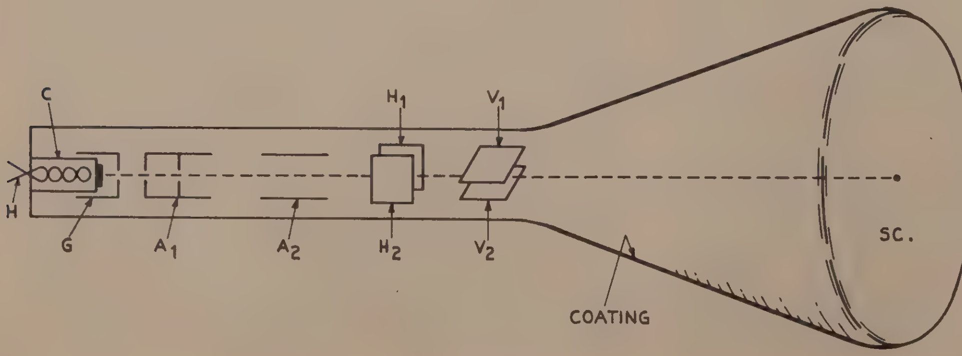


FIG. 10-1.—The elements of a cathode-ray tube: *H* is the heater; *C* the cathode; *G* the control grid; A_1 and A_2, the first and second anodes respectively; H_1, H_2 and V_1, V_2, the horizontal and vertical deflection plates, respectively; and *SC*, the screen.

source may be sufficiently intense, any adjustment, for example some setting of the stop S may reduce the amount of light conveyed from L to lens O, will interfere with the intensity of the spot; and if this adjustment is such that the light transferred is not sufficient, the intensity level of the spot upon the screen will not be suitable. Bear in mind that at this time we are concerned solely with the intensity or the brightness of the spot or image.

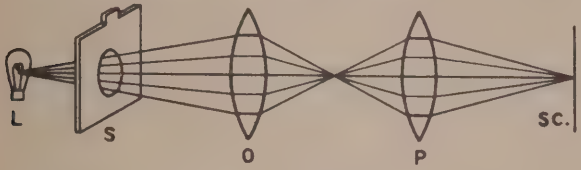


FIG. 10-2.—An optical system which is analogous to the elements of a cathode-ray tube which bring the electron stream to a focus on the screen.

Two variables are directly associated with the intensity or luminosity of the spot. If we desire to vary the intensity of the light, we have two variable controls, either or both of which may be used. If we should decide to keep the brilliancy of the lamp fixed, adjustment of the opening of the stop S would give us the desired control. On the other hand, if we desire to keep the adjustment of the stop opening fixed, the filament control in the lamp circuit would give us the intensity control. There is no reason for a dual control of this type, so we can forget for the moment any variation in the brilliancy of the lamp and use only the stop opening as our intensity control. When the stop is opened wide, the full intensity is realized, because the maximum amount of light goes through. When the stop opening is closed, no light passes through. Any setting of the stop opening, between wide open and fully closed, will provide a light intensity varying between the two limits.

For any one particular distance between the light source and the screen, there is one setting of the lens, which will result in the most sharply defined area of light. This operation or adjustment of the lens to most sharply define the rays of light upon the screen is known as focusing.

In connection with the optical analogy offered to illustrate focusing in the television picture tube, we are concerned with light rays which converge to a pin point. The adjustment of lenses O and P constitutes focusing with two variables. If lens O is fixed in position, focusing can be accomplished by moving lens P. On the other hand, if lens P, is fixed, all the focusing can be done with lens O. With lens P fixed in position, adjustment of stop S and lens O will result in an optimum setting, which will provide greatest intensity and greatest definition; in other words, a spot of the proper intensity is now properly focused upon the screen. Incorrect adjustment of stop S with correct adjustment of lens O will result in a properly focused spot of light upon the screen, but one of insufficient brilliance. On the other hand, correct adjustment of the stop and incorrect adjustment of lens O will result in light upon the screen, but not of the correct intensity or focus. The intensity also suffers when the light is out of focus because proper concentration does not take place. Instead of having all the light concentrated into one spot, the rays picked up by the lens are diffused over a greater portion of the screen, with reduced light at any one point.

It is significant to note that with the position of the light source L, the opening of stop S, and the screen position permanently fixed, the projection of the best defined spot is determined by the relative positions of lenses O and P. Such focusing can also be done with a single lens in the place of O and P.

Electrostatic Focusing

Let us now see how closely the optical analogy of focusing holds for the electronic method employed in the television picture tube. Referring to Fig. 10-1, H is the heater which raises the cathode temperature to the proper operating point for the emission of electrons. C is the cathode, which emits the electrons and is equivalent to the light source in Fig. 10-2. G is the control grid and is equivalent to the light gate, or stop. A_1 and A_2 are the first and second anodes which are operated at high positive potentials. These electrodes operating in unison are called an electron gun because they concentrate the electrons into a narrow beam and direct this beam against the screen with sufficient force to cause the required degree of fluorescence.

The control grid differs from the conventional grid in an ordinary vacuum tube, being a small hollow cylinder. The end of the cylinder toward the cathode is completely enclosed except for a small pinhole through which the electrons pass. Just as stop S in Fig. 10-2 serves as a light gate, the control grid functions as an electron gate or stop because of the following. The electrons which are emitted from the cathode are negative particles. For any one temperature of the cathode, the maximum number of electrons are emitted for that temperature. If we desire to control the intensity of the spot, we must have some means of controlling the number of electrons allowed to pass into what finally becomes the beam. The intensity of the spot upon the viewing screen is dependent upon the number of electrons in the beam, just as the intensity of the spot in the optical analogy is dependent upon the amount of light allowed to pass to the lenses.

Control of the number of electrons which are permitted to pass through the electron gate is accomplished somewhat in the manner employed in the regular vacuum tube, that is, by the application of a negative bias to the control grid. The negative bias applied to the grid repels the electrons emitted from the cathode and definitely controls the number which pass through the pinhole aperture. Reducing the bias naturally allows a greater number of electrons to pass through aperture and thus pass under the influence of the adjacent electrode. A_1. The greater the number of electrons permitted to pass through the control grid, the greater the intensity of the spot which appears upon the viewing screen.

The electrons emitted from the cathode have a certain velocity but not the velocity required. Therefore, a positive voltage is applied to A_1 to attract sufficient quantities of electrons and to accelerate their movement through the electron gate *G* towards the viewing screen. The pinhole aperture and the negative bias on the control grid plus the attracting influence of anode A_1 combine to produce the initial converging force, for the diameter of the electron beam passing through the control grid is much less than the beam diameter at the cathode.

The control grid electrode serves another purpose in addition to controlling the number of electrons permitted to pass through the aperture. As a result of the negative charge applied, this grid limits electron emission from the outer portions of the cathode. Only the center portion of the cathode, parallel to the aperture, emits freely. Of course, the exact extent of this condition depends upon the bias adjustment. The control grid, in conjunction with the cathode and the first anode A_1 which applies the initial acceleration force to the emitted electrons, constitute an electronic lens.

From what has been said, it is clearly evident that the primary control of the electron beam is the intensity control, that is, the control grid electrode bias adjustment. When the bias adjustment of the control grid is such that no electrons, or insufficient electrons, pass through the electron gate aperture, no electron beam will be formed. On the other hand, if sufficient electrons are made to pass through the electron gate, a spot will appear upon the screen irrespective of the adjustment of the anode A_1 voltage. This corresponds to the case of the simple optical system shown in Fig. 10-2, where, if light is passed through the stop, it will be visible upon the screen even if the adjustment of lens *O* is not correct.

The voltage applied to anode A_1 is the primary focusing voltage, since the first focusing action takes place in the region of the first anode. Anode A_1 is a cylinder having a small aperture in the end nearest to A_2. The initial beam of electrons is converged in A_1 for passage through this aperture. Those electrons which are too divergent from the beam do not pass through the aperture. A comparison between the simple optical system and the electrical optical system in the picture tube shows a parallel between *L*, *S*, and *O* in Fig. 10-2 and *C*, *G*, and A_1 in Fig. 10-1.

The main focusing action in the picture tube takes place between anodes A_1 and A_2. A fixed high voltage is applied to anode A_2. This voltage is positive with respect to the cathode and much greater in value than the voltage applied to anode A_1. This voltage on A_2 accelerates the electrons to the velocity required to produce fluorescence when they strike the screen. The electrons which pass through the beam-defining aperture in A_1 are brought to a focus as a result of the combined action of A_1 and A_2, but at this time they are caused to converge into a spot at a specific distance from A_2, namely, upon the screen. Anode A_2 also has a beam-defining aperture, so that electrons which are too divergent from the main beam are not permitted to pass through.

During the discussion of the optical system, lens *P* was fixed and focusing was accomplished by adjustment of lens *O*. With the position of *L* and the adjustment of *S* also fixed, it is obvious that focusing of the light upon the screen *Sc* is governed by the relative positions of the two lenses, *O* and *P*. A corresponding condition is true in the picture tube, but in this case, with cathode emission fixed, the focusing of the beam upon the screen is determined by the ratio of voltages applied to anodes A_1 and A_2. With a fixed voltage on A_2 the proper focusing adjustment is made by varying the voltage applied to anode A_1.

We made the statement that if the intensity adjustment (the bias applied to the control grid) was correct, electrons would strike the viewing screen regardless of the setting of the focus control which varies the voltage applied to A_1. It is also important to remember that if the intensity adjustment is such that insufficient electrons are passed, or if the emission from the cathode is insufficient, variation of the focusing voltage may cause the spot to disappear from the screen. There are several possible reasons for this. One is that as a result of incorrect focusing, insufficient acceleration is given to the electrons and comparatively few reach the high voltage anode A_2. Another reason is the diffusion of the beam, which results in less than the required amount of light appearing on the screen.

The production of a bright and sharply defined spot upon the screen is accomplished by correctly controlling the intensity and the focusing of the electron beam. The size of the spot is influenced by both the intensity and focusing controls.

Magnetic Focusing

In our discussion on focusing the electron beam in the picture tube, we have described the method using electrostatic fields. However, the same results can be obtained by means of magnetic fields. This reference to magnetic focusing must not be confused with the application of magnetic fields for the purpose of deflecting the beam in the formation of the image. This reference relates to the proper concentration of the electron beam in order to provide a small, sharply defined spot on the screen.

Most of you will recall from your study of elementary electrical theory, that a circular magnetic field is established around a conductor carrying an electric current. By placing a current-carrying coil around the neck of the picture tube near the location of anode A_2, the magnetic field can be made to act upon the beam of electrons within the tube. The effect of this action is to force divergent electrons to assume a helical movement which terminates in a sharply defined spot on the screen when the beam is focused properly. This helical movement is the result of two forces. The first force is a property of the electrons diverging from the electron beam which is being accelerated by the positive voltages on anodes A_1 and A_2. The second force is that of the magnetic field which acts at right angles to the first force. As long as the electron beam travels along the axis of the picture tube, it is not affected by the lines of force of the magnetic field. However, when the beam passes out of the converging influence of the positive anode, electrons on the surface of the beam tend to spread out. It is these electrons that come under the influence of the magnetic field which throws them into a helical movement in the effort to force them back into the original axis of the beam.

The focusing coil is connected in series with the low-voltage power circuit, and the current through the coil is controlled by a rheostat. Focusing is accomplished by varying the amount of current flowing through the coil until the point of optimum focus is reached. This magnetic method of focusing has the advantage of reducing certain detrimental effects which occur on the screen of the tube. These detrimental effects caused by a condition known as ion spot will be covered in more detail later.

In some instances the internal structure of the picture tube may be considerably different from the tube shown in Fig. 10-1. Some types of picture tubes have a graphite coating on the inner surface beginning near the fluorescent screen and extending back to anode A_2. This conductive coating is in some cases connected internally to the second anode and in other cases is brought out to a separate terminal located on the side of the tube. A positive voltage equal to, and at times greater than, the voltage on A_2 is applied to this coating. The coating serves as a return circuit for the electrons that bounce back after contacting the screen, and also as an additional anode providing greater accelerating force to the electron beam.

Quite often where magnetic focusing is employed, anode A_2 is omitted entirely in the fabrication of the tube, and this coating, which is then brought out to a separate terminal, serves as anode A_2. The focusing coil is positioned in the neck of the tube at the point where the electron beam normally comes under the influence of anode A_2.

Deflection of the Beam

There are two very important functions performed in the picture tube. The first is the focusing of the electron stream into a narrow concentrated beam of electrons. The second is the subjection of this beam to the influence of an electrostatic or magnetic, or combination of both, field in order to secure the proper deflection of the beam. Deflection of the beam results in a scanning pattern, or raster, on the screen of the picture tube.

We made the statement that two types of deflection are available, namely electrostatic and magnetic. Electrostatic deflection is employed in the smaller picture tubes (5 to 7 inches), but because of the extremely high deflecting voltages that would be required, magnetic deflection is used in tubes having larger diameters. Some of the factors to be considered in deflection are the velocity of the electron beam and the intensity of the electrostatic or magnetic fields employed. Also, it must be remembered that electrons are particles bearing a negative charge and as such are repulsed by a negative field or attracted by a positive one. In addition, electrons in motion are surrounded by a magnetic field in the same manner as a current-carrying conductor, and can be affected by external magnetic forces.

In the discussion of the deflection of the electron beam, we will consider first deflection with an electrostatic field and then with magnetic fields. As far as deflection is concerned, the nature of the electron emitter and the type of accelerating anode or anodes used are of no consequence. The primary consideration is the existence of the electron stream within the deflecting field.

Electrostatic Deflection

Electrostatic deflection is accomplished by the application of voltages to the deflecting plates in the picture tube. These plates are identified in Fig. 10-1 as H_1, H_2, V_1, and V_2. The position of the electron beam without any voltages applied to these plates is midway between them; this places the spot in the center of the viewing screen. This position of the spot midway between the plates also exists if no potential difference exists between the plates. In other words, if both plates of each pair are at the same potential, the effect upon the beam is as though there were no voltage whatsoever applied to either pair of plates.

To appreciate properly what happens in the picture tube when a potential difference exists between the plates of each pair of deflecting plates, it is necessary to consider each pair separately. Particular reference was made to the fact that the electron stream or beam consists of a group of moving negative charges; also, that these charges are repelled by similar negative charges or by a negative field and that they are attracted by positive charges or a positive field. Suppose then, that the beam of electrons is moving between the two deflecting plates as shown in Fig. 10-3. By means of connections to these plates, it is possible to apply a positive voltage to the upper plate V_1 and a negative voltage to the lower plate V_2. With such connections, the positive charge upon V_1 will attract the stream of electrons towards it, whereas the negative charge upon the lower plate V_2 will repel the stream, so that the entire beam will be deflected toward the upper plate V_1.

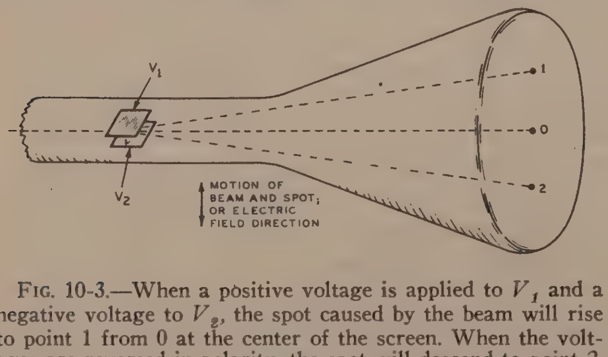


FIG. 10-3.—When a positive voltage is applied to V_1 and a negative voltage to V_2, the spot caused by the beam will rise to point 1 from 0 at the center of the screen. When the voltages are reversed in polarity, the spot will descend to point 2.

The normal direction of the beam is as shown by the middle dotted line in Fig. 10-3 and the position of the spot upon the screen is O. When the difference of potential previously described exists between the plates, the beam is pulled out of line as it passes between the plates and the spot upon the screen is moved to the point indicated by the number 1, the beam being shown as a dotted line. The extent that the spot is moved up away from its normal position is determined by the magnitude of the positive charge upon this plate, with respect to the lower plate.

The limitation of the movement of the spot, or deflection of the beam, with a certain potential difference between the two plates, is a result of the voltages applied to the accelerating anodes. These voltages project the electrons in a beam toward the screen. Since the velocity of these electrons in motion is very great, a certain force is necessary to deflect them or pull them away from their normal line of travel. The potential difference between the deflecting plates is the force which tends to pull the stream out of its normal axis of travel. The higher the accelerating voltage which starts the beam in motion, the greater the force required to pull it out of line. It should be understood that there are also other factors present in the tube which may offset the effect of the high accelerating voltage, so that two tubes of unlike accelerating voltages may require a like value of deflecting voltage in order to move the beam the same distance. The degree of sensitivity to deflection is known as the deflection factor of the tube and varies with different tubes.

If the voltage to the two deflecting plates is reversed as shown in Fig. 10-3, so that V_1 is negative and V_2 is positive, the beam and the spot shift in a direction opposite to that which took place before. The basis of this shift is exactly as stated before, except that the polarity of the deflecting plates has been changed. Now the electron beam is attracted by V_2 and is repelled by V_1. Since the hypothetical voltage applied to the two deflecting plates is the same value as before, the potential difference between the plates is the same and the spot moves the same distance in the opposite direction to point 2.

If a device is used whereby the voltage applied to the two plates is automatically reversed at a very slow rate, you would see the beam shift from position O to position 1, back through position O to position 2, thus tracing a line of light on the screen.

At this time we want to call to your attention one very significant fact. The electric field existing between these plates is like the electric field existing between the plates of a two-plate capacitor and a specific amount of capacitance does exist between a pair of deflecting plates. The electric field exists perpendicular to the plane of the plates, so that the plates which are horizontal with respect to the front of the tube cause the vertical deflection. The motion of the beam is parallel with the electric field. The term "vertical deflection plates" refers to the motion given to the beam by those plates, with respect to the normal position of the tube, and not to the orientation of the plates inside the tube. This applies to the term "horizontal deflection plates" as well.

The action we have described in regard to the vertical deflecting plates also occurs when a potential difference exists between the horizontal deflection plates, H_1 and H_2 in Fig. 10-4. With H_1 positive with respect to H_2, the spot will move from O to position 1. If the polarity of the charges is reversed, so that H_2 attracts the stream, the beam is pulled in the opposite direction and the spot moves to position 2.

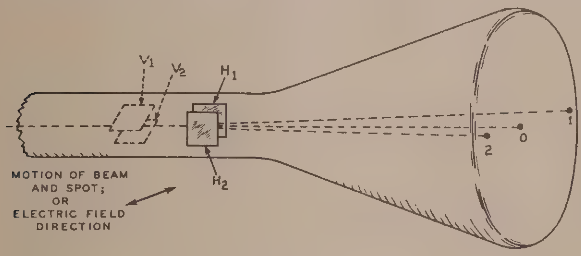


FIG. 10-4.—If horizontal deflection plate H_1 be made positive and H_2 negative, the spot will move to the right from the center of the screen 0 to point 1. When the polarity of the voltages is reversed, the spot will move toward point 2.

This is basically what happens in the tube when electrostatic deflection is employed. With this understanding, it is not difficult to visualize what occurs within the tube when the sweep voltages described in chapter 9 are applied to the deflecting plates.

Magnetic Deflection

As in the case of magnetic focusing previously discussed, the electron beam is subjected to the force of strong magnetic fields in order to deflect it across the screen of the picture tube. In most tubes using magnetic deflection, the two pairs of deflecting plates are left out of the tube, and a magnetic deflecting yoke is placed around the neck of the tube at approximately the position occupied by the deflecting plates.

This magnetic yoke consists of two pairs of deflecting coils mounted at right angles to each other, in the manner of the deflecting plates used for electrostatic deflection. The two coils comprising each pair are connected in series and are mounted opposite to each other on the neck of the tube. When current is passed through the coils, the magnetic field which is created bends the electron beam in the same manner as the electrostatic field previously discussed.

In operation, deflecting currents from the sweep circuits are passed through the coils, causing the electron beam in the tube to trace out the raster on the face of the tube. The deflecting yoke must be adjusted in position on the neck of the tube so that the picture will appear in correct position for the viewing audience.

Combined Methods of Focusing and Deflection

Quite often combinations of electrostatic focusing and magnetic deflection and vice versa will be encountered. It has been stated that on the larger picture tubes, magnetic deflection is used. One reason for this is that magnetic deflection provides uniformity of focus at all points on the viewing screen, which is not true in the case of electrostatic deflection. Another reason is that the hazard of electrical shock is reduced since the positioning controls, which are normally connected to the deflection plates in electrostatic deflection circuits, are no longer in the high-voltage circuit.

If electrostatic focusing is used, electrostatic deflection should also be used to minimize defects to the screen of the tube caused by ion bombardment. Magnetic focusing and electrostatic deflection is not entirely satisfactory because of the nonuniformity of focus at all points on the screen, as was mentioned previously.

Magnetic focusing reduces the effect of ion bombardment, and magnetic deflection provides a more uniform focus over the area of the screen. From this it can be seen that magnetic focusing combined with magnetic deflection provides a better television picture than does either of the combined methods. This last is the favored method; however, certain design problems in particular sets will determine which method is incorporated in the television receiver.

ION SPOT

After a picture tube which employs electrostatic focusing and magnetic deflecting methods has been in operation for a time, a small round spot may appear on the screen of the tube. This spot constitutes a blemish in the screen material and impairs the quality of the television picture. The cause of this spot has been traced to negatively charged heavy particles in the electron beam which strike the screen with great force, causing eventual deterioration of the screen material. These heavy particles are negative ions and the blemish on the screen, which appears dark brown in color, is called an ion spot.

When the cathode of the picture tube reaches operating temperature, electrons and negative ions are emitted from the cathode material. Of course, all of the negative ions may not be emitted from the cathode material, some may be formed by other means. However, all of these negative particles are formed into a concentrated beam as a result of the electronic lens system within the tube. The high accelerating voltage on the second anode increases the velocity of the beam to a high value. The action of the magnetic fields of the deflecting coils easily deflects the electrons in the beam, but because of their relatively greater mass, the ions are not deflected and continue to travel straight along the axis of the tube and strike the center of the screen at high velocities.

It can readily be seen that if this ion action were allowed to continue, rapid deterioration of the screen material, with attendant short life span for the tube, would result. Since picture tubes are rather expensive items, it becomes necessary to eliminate the detrimental action of the ion spot if prolonged and efficient service is desired.

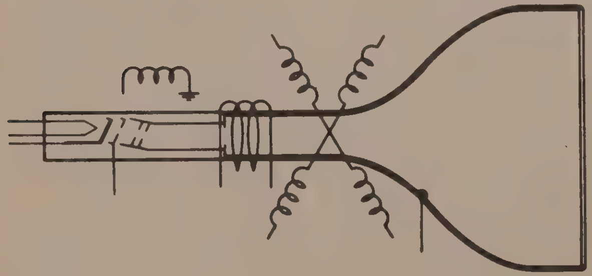


FIG. 10-5.—In order to get rid of the negative ions in the electron stream, the "gun" is placed at an angle so that the ions are aimed at one side of the anode while a bending coil causes the ion-free electrons to turn and go along the axis of the tube.

The problem involves either the removal of these heavy negative particles from the electron beam by means of a trap method which would reject the ions while permitting the electrons to pass freely, or, in some other way, to prevent the ions in the beam from striking the fluorescent screen. As a matter of fact, both of these methods are utilized on occasion and have been very effective in eliminating the ion spot. These two methods, both of which involved a change in tube structure methods, will be described in detail in the following discussion.

Ion Trap Methods

One of the methods developed to eliminate ion spot from the screen of the tube is the bent gun method. The electrodes within the tube which constitute the electron gun are bent so that the path of the electrons is toward the side of the anode as shown in Fig. 10-5. However, by means of the magnetic field from a "bending" coil placed around the neck of the tube, the electrons are bent back into line so that their path lies along the axis of the tube. The heavy ions are not affected by the magnetic field and, therefore, travel straight along the original path and strike the side of the anode, or, in some cases, a disk which serves to collect the ions. In this manner, the undesired negative ions are eliminated entirely from the electron beam. Following this, the stream of ion-free electrons is then focused, accelerated, and deflected in the usual manner.

In chapter 1 mention was made of the practice of coating the inner surface of the luminescent screen material in the picture tube with a thin backing of aluminum film in order to increase the brightness of the images on the face of the tube. In addition, the application of this metallic film provides another method of correcting the condition caused by the ion spot on the screen. It has been discovered that a film of aluminum coated over the fluorescent material will readily pass the electron beam but will completely stop the heavy ions, thus preventing them from bombarding the screen material. Of course, a number of factors must be taken into consideration, such as the thickness of the metallic film, the gas content of the tube, and the operating potentials applied to the high-voltage anode. However, prolonged tests have proved that tubes having a metallic backing of the proper thickness offer no restrictions to passage of the electron beam and yet do not develop an ion spot when the applied operating voltages are normal.

PROJECTION TUBES

All of the tubes discussed up to this point have been of the direct-viewing type, that is, the television picture appears on the face of the tube and is viewed directly by the observers. One of the complaints about the presentation of television pictures in earlier receivers was the small size of the reproduced image, since the tubes then available were usually not larger than 7 inches in diameter. In an attempt to satisfy the demand for a larger picture, the television industry turned to the development of receivers using picture tubes with larger diameters. However, this offers only a partial solution, for as the size of the tube is increased, the production costs and the costs of storage and transportation increase rapidly. These tubes are of the very high-vacuum type and must be fabricated to withstand extremely high air pressures, with the attendant hazard of injury by implosion or breakage of the tube. Sufficiently large pictures for large coverage cannot be attained even with use of the 20-inch picture tube, which is the largest tube in use in commercial television receivers today.

The most practical solution from all viewpoints seems to be the projection of the television picture upon a viewing screen of the desired size, similar to the manner in which motion pictures are projected. In this method, the image on the face of the tube is directed through an optical system consisting of correcting lenses and reflecting mirrors and then projected to the screen. This permits the use of a small size tube, as the original picture can be greatly enlarged in the optical system. However, another problem arises, and that is the creation of a sufficiently bright image on the face of the tube. There is a loss in light intensity when light is passed through an optical system. This must be taken into account when designing such a system for projection television. A limit is thus quite obviously placed upon the minimum size of the picture tube.

We know from our discussion of the electron beam that the intensity of light produced on the flourescent screen is directly proportional to the velocity of the beam at the point of impact. The obvious answer to the question of screen brightness would seem to be an increase in the acceleration of the electron beam, and this is exactly what is done. For home receivers, accelerating voltages as high as 30,000 volts are applied to the accelerating anode; while for theater projection, voltages as high as 80,000 volts are used.

In projection-type television receivers used in the home, a sufficiently bright, sharply defined picture as large as 12 by 16 inches can be obtained from a picture tube having a diameter no greater than 2½ inches. In theater projection of television pictures an 18-by-24-foot picture can be obtained from a 15-inch tube.

It is not our purpose in this discussion of picture tubes to become involved in a description of the various projection methods resorted to in order to obtain a picture of the desired size, brightness, and sharpness of detail. However, it seems that a brief reference to one of the most popular systems in use at the present time will illustrate more graphically how the television tube functions in such a system.

Fig. 10-6 illustrates this projection system which is an adaptation of the Schmidt principle developed originally for use in telescopes used by astronomers. Referring to the figure, the image on the face of the tube is directed onto the spherical mirror in the bottom of the cabinet. From there the image is projected through the correcting lens which corrects the aberration caused when the light strikes the spherical mirror. The light then strikes the inclined mirror at the top of the cabinet and the pic-

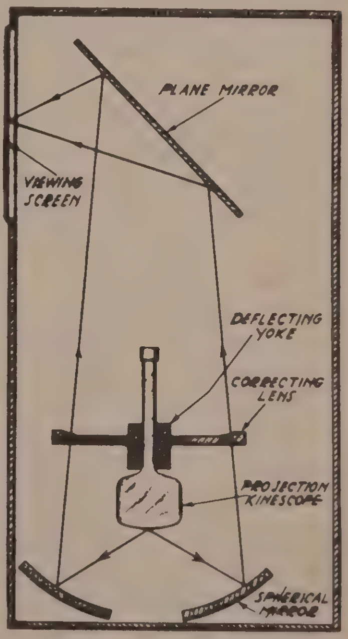


Fig. 10-6.—The optical system of a projection television receiver, which employs the principles originated by Schmidt for use in telescopes. The image of the picture on the screen of the picture tube is reflected upward by the spherical mirror going through the correcting lens into the inclined mirror. From here it is reflected onto the viewing screen.

Courtesy RCA

ture is reflected onto the viewing screen in enlarged size. Modifications of this system in which the light is "folded" back once or twice has resulted in a greater utilization of the available light with reduced losses.

The most currently popular projection tube used in commercial television receivers is the 5TP4 projection Kinescope. This is a 5-inch tube having electrostatic focusing and magnetic deflection. The accelerating voltage applied to the second anode is 27,000 volts (maximum) with a maximum of 6,000 volts being applied to anode number one. The required grid drive voltage is about 40 volts.

The tube has an internal coating brought out to a terminal on the side of the tube which serves as anode one. An external insulating coating, which is grounded, in operation forms a capacitance with anode two in the order of 500 $\mu\mu f$. This capacitance serves as a filter capacitor to the high voltage applied to anode two. The curved screen is coated with phosphor which fluoresces white and has a medium persistency. The tube can be mounted in any position and is ideally suited for employment in the Schmidt projection system previously described.

A recent development in a small projection tube is the MW-6 (nomenclature for experimental tube) which has been developed by the North American Philips Company. This tube has a 2.5-inch diameter and an over-all length of 10.5 inches. The image reproduced on the face of the tube measures 1.4 by 1.86 inches and from this image, a 12-by-16-inch picture with satisfactory brightness, resolution, and contrast is projected onto the viewing screen. The tube utilizes magnetic focusing, magnetic deflection, and a moderate deflection angle. It features a small neck with an accurately positioned electron gun. The accelerating voltage applied to the second anode through a special type terminal brought out to the side of the tube is 25,000 volts. By proper design of the electron gun a very small diameter spot (0.003 inch) is produced on the screen of the tube, permitting adequate resolution of the small size image appearing on the tube face. The focusing and deflecting requirements are the same as for the 10BP4 direct-viewing picture tube operated at 9,000 volts, which permits the substitution of this projection tube in place of the direct-viewing 10BP4.

The second anode is coated on the inside of the tube, and a glass cup is sealed to the tube around the second anode terminal to insure corona-free operation. A second coating of Aquadag on the outside of the tube is grounded and used as a static shield. The two coatings form a capacitor of approximately 300. $\mu\mu f$ which provides final smoothing action of the high voltage applied to the tube. The average beam current is 90 microamperes and on highlight peaks it reaches the value of 500 microamperes. The required grid drive voltage is about 50 volts peak.

This tube is employed in a projection system using "folded" Schmidt optical principles in a triangular arrangement. Because the tube itself is an element in the optical system, the face plate of the tube must be optically correct. This face plate is molded of special glass which will be not discolored by the low intensity, soft X radiation which is produced by the 25,000-volt electron-beam bombardment. A phosphor screen with an aluminized backing is used which increases the light output, improves over-all contrast, and eliminates the necessity for an ion trap. The tube is powered from a special 25,000-volt pulse-type voltage-tripling power supply which uses a special control circuit for improved voltage regulation.

This new projection tube with its optical system and power supply is incorporated in a special kit of small size and is intended for installation in television receivers having the power supply requirements of a 10-inch direct-viewing picture tube such as the 10BP4. This permits the conversion of existing receivers, using direct-viewing 10-inch tubes with the proper voltage requirements, into projection sets giving larger and more satisfactory group coverage.

Various sizes of projection tubes are used in receivers designed for television theater projection, but these tubes function basically in the same manner as the picture tubes described here. The important thing to remember in the operaion of any picture tube is the formation of a sharply focused electron beam and its deflection across the screen of the tube. These functions are the same regardless of the shape and size of the tube, any modifications in the structure of the tube, and the application to which the tube is finally put.

CHAPTER 11

POWER SUPPLIES

By Henry Chanes

As a general rule, two separate power supplies are used to supply the d-c voltages for television receivers. A low-voltage supply provides approximately 300 volts for the tubes associated with the amplifier, deflection, and sync circuits of the receiver. A separate high-voltage power supply provides the high voltages required for the picture tube. These voltages vary from 2,000 to 10,000 volts for direct-viewing tubes and as high as 30,000 volts for projection tubes.

Low-Voltage Power Supplies

The low-voltage power supplies used in television receivers are very similar to those used in radio receivers. However, the regulation in a television low-voltage power supply is generally better than that in a radio receiver because of the importance of preventing variations in the numerous circuits from interfering with each other and with the picture. The hum level is also kept down to a lower value because of the large number of circuits and stages, and because hum which would be inaudible in a sound receiver may cause appreciable distortion of the picture.

The low-voltage power supply used in the RCA model 621TS receiver is shown in Fig. 11-1. The transformer *T*106 is a conventional power transformer with sufficient capacity to supply the plate and heater currents required by the receiver. To prevent any undesirable coupling between the picture tube and the other circuits in the receiver, there is a separate heater winding for the picture tube. The transformer also has a separate 5-volt winding for the heater of the 5V4-G damper tube to prevent any interaction between the damping circuit and the power supply. The capacitors *C*143 and *C*144 across the primary are the usual filter capacitors to prevent any noise on the power line from coming through the receiver. The filtering in this power supply is quite extensive as can be noticed from the large number and high value of the filter capacitors used. This filtering is important not only to keep the hum down to a very low level, but also to prevent any coupling between the many different circuits supplied by this one power supply.

The voltage drop across potentiometers *R*166 and *R*152 are used to control the horizontal and vertical centering respectively. Direct current is made to

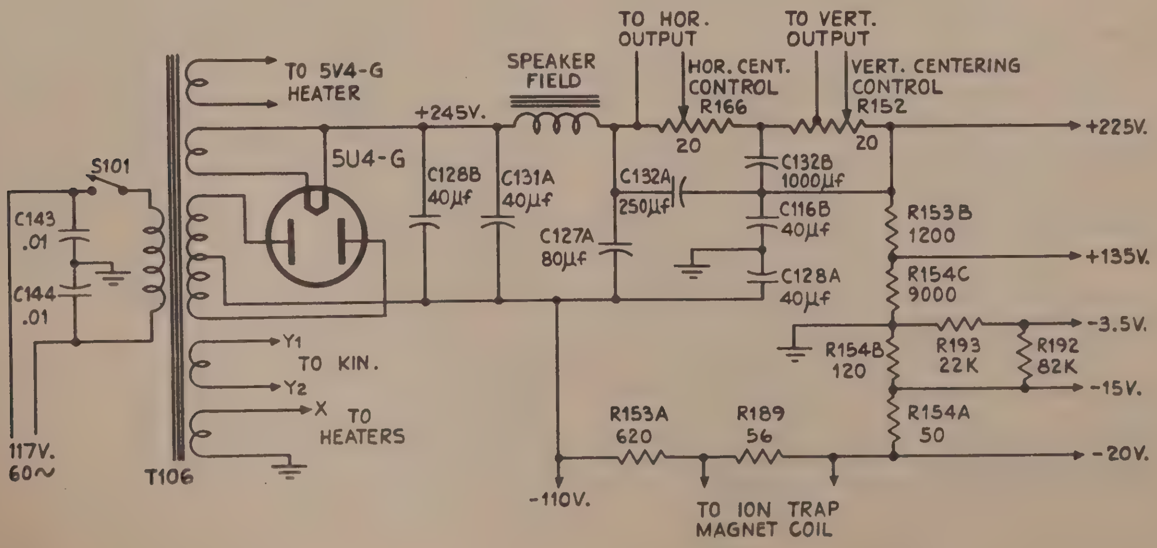

TO HOR. OUTPUT
TO VERT. OUTPUT
TO 5V4-G HEATER
SPEAKER FIELD
HOR. CENT. CONTROL R166
VERT. CENTERING CONTROL R152
+245V.
20
20
+225V.
S101
5U4-G
C128B 40µf
C131A 40µf
C132A 250µf
C132B 1000µf
C116B 40µf
C127A 80µf
C143 .01
C144 .01
R153B 1200
+135V.
R154C 9000
C128A 40µf
-3.5V.
R154B 120
R193 22K
R192 82K
-15V.
Y1
TO KIN.
Y2
X
TO HEATERS
R153A 620
R189 56
R154A 50
-20V.
117V. 60~
T106
-110V.
TO ION TRAP MAGNET COIL

Fig. 11-1.—The low-voltage power supply circuit used in the RCA model 621TS receiver.

flow through the yoke in addition to the saw-tooth sweep current. The saw-tooth current makes the spot sweep back and forth, or scan, the face of the tube, but d.c. merely makes the spot move to one side or the other and stay there as long as the current flows. The way the spot moves depends on the polarity of the current. The voltage between the arm and the center tap of the vertical centering control *R*152 can be made either positive or negative, depending upon which side of the center tap the arm is placed. This voltage will cause d.c. of either polarity to flow in the vertical yoke and enable the picture to be moved up or down

In the horizontal centering control circuit, the situation is somewhat different. The damping tube in the horizontal sweep circuit will, because of its rectifier action, produce d.c. The action of this tube is discussed in more detail in chapter 9 which covers sweep circuits. Because of this d.c. produced by the damping tube, it is necessary to produce d.c. of opposite polarity in order to center the picture. This is done by using the voltage drop between the arm and one side of the horizontal centering control *R*166, since d.c. of a constant polarity is required by this circuit.

The string of resistors following the centering controls make up a bleeder to supply the different positive and negative voltages required by the different circuits in the receiver. The voltages shown at the different points in the bleeder circuit are determined by the bleeder current and the currents drawn by the load on each output of the power supply. The power supply is grounded only at the junction of the resistors *R*154*C* and *R*154*B*, so that negative as well as positive voltages with respect to ground are available.

The low-voltage power supply in the Consolidated Television model 2315 consists of a single heater supply plus three separate d-c supplies that use selenium rectifiers instead of vacuum tubes. The use of selenium rectifiers in television receivers has several advantages over the conventional power supplies.

1. By the use of voltage-multiplying circuits where higher voltages than the power lines are necessary, the power transformer can be eliminated. This appreciably cuts down the weight and size of the receiver. Also, the danger of stray fields from the transformer affecting the video circuits and producing interference in the picture is removed.

2. Separate supplies for different circuits in the receiver are practical with the use of selenium rectifiers. This eliminates the large bleeder resistors that are necessary when a common supply is used for all the circuits, resulting in greater efficiency and lower operating temperature. Undesirable coupling between different circuits of a receiver due to a common power supply also is reduced by the use of more than one supply.

3. The selenium rectifier does not require heater current. It is small, light, and easy to install. It runs cooler than a tube and generally has a much longer life.

The operation of the selenium rectifier is easily understood. Its rectifying action depends on a property often found at a junction of two dissimilar metals, i.e., that electrons flow more readily in one direction than in the opposite direction.

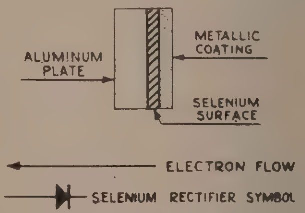


FIG. 11-2, right.—The electron flow through a selenium rectifier is from the metallic coating-selenium surface toward the aluminum plate, as indicated by the arrow.

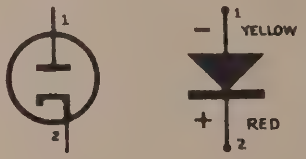


FIG. 11-3, left.—In the schematic symbol for the selenium rectifier on the right, the electron flow is from 2 (identified by a red dot) upward to 1 (identified by a yellow or no dot).

A typical unit found in present-day television receivers consists of several small plates, usually square in form, stacked together at their centers, and utilizing two lugs for connection to the circuit. Each element is simply a supporting aluminum plate, coated on one side with selenium, and then having a metallic coating over the selenium. This selenium "sandwich" ensures adequate electric contact with both sides of the selenium. By proper heat processing, a rectifying film is formed between the selenium and the metallic coating. The metallic coating has a great number of free electrons that can flow through the selenium, and into the aluminum plate. However, the selenium has very few free electrons available, so that the electron flow in the opposite direction is very limited.

If the negative terminal of a power source is connected to the metallic coating, the flow of current toward it is very small, for we have seen that there are few free electrons available. Consequently, the electron flow in Fig. 11-2 is from right to left. The symbol used to represent the selenium rectifier and similar types, shown in Fig. 11-2, was adopted before the electron-flow theory was accepted, but has never been revised. Therefore, when a selenium rectifier replaces a diode rectifier in a circuit diagram, the corresponding terminals are as indicated in Fig. 11-3.

A typical type of selenium rectifier is the Federal Type 403D2625, which has a maximum d-c output of 100 ma and a drop of 5 volts across it. The two lugs may be marked positive and negative, or they may be color coded; positive is indicated by a red dot, negative by a yellow dot or blank, as in Fig. 11-3. Care must be taken that the proper polarity connections are made, remembering that B+ is obtained from the *cathode* of the rectifier tube, and so the positive lug (red dot) must be where the cathode was formerly connected.

The low-voltage power supply of the Consolidated Television model 2315 is shown in Fig. 11-4. The purpose of the heater transformer is obvious, so let us look at the first of the selenium rectifier supplies. This one has an output of 440 volts. To get this high voltage without the use of a step-up transformer, a voltage-tripling circuit is used. This circuit will theoretically have an output of 3 times the peak value of the input voltage. If we assume the line voltage to be 115 volts rms, the peak value is 115 × 1.4 or approximately 160 volts. The output of the voltage tripler then should be 480 volts. Actually this circuit delivers 440 volts. The loss in voltage from the ideal value is due to the fact that current is being drawn by the load on the supply and the drop in the rectifiers themselves. There is also some voltage drop in the current-limiting resistors *R*65 and *R*66.

When the side of the line that is connected to *R*66 goes positive, the rectifier *RX*-1 will conduct and charge *C*61 to the peak value of the line voltage. Let us call this voltage E_m. The polarity of the charge on *C*61 is shown in Fig. 11-4. On the next half cycle *RX*-1 will not conduct, but capacitor *C*61 will act in series with the line and, with the rectifier *RX*-2 now conducting, will charge the capacitor *C*60 *to* $2E_m$ with the polarity shown in Fig. 11-4. On the next positive half cycle, *C*60 will act in series with the line and a total voltage of $3E_m$ will be applied to the rectifier *RX*-3 which is now conducting. Capacitor *C*62 will be charged up to $3E_m$. This is one of the output voltages. The other is taken from the other side of the filter which consists of *C*62, *R*82, and *C*63, and due to the voltage drop in *R*82, is equal to 380 volts.

The second selenium rectifier supply delivers 105 volts. This section contains the limiting resistor *R*67, the selenium rectifier *RX*-4, and the filter consisting of *C*64, *R*82*A*, and *C*65. The rectifiers used in this circuit are the same type that are used in the voltage tripler supply. They are Federal Type 404-D2795 and have a d-c capacity of 200 milliamperes.

The third section of the power supply is a bias supply. It has an output of —105 volts. Notice that the selenium rectifier *RX*-5 is connected with the terminals reversed in order to obtain an output which is negative with respect to ground. The filter capacitors have their positive terminals grounded for the same reason. The selenium rectifier *RX*-5

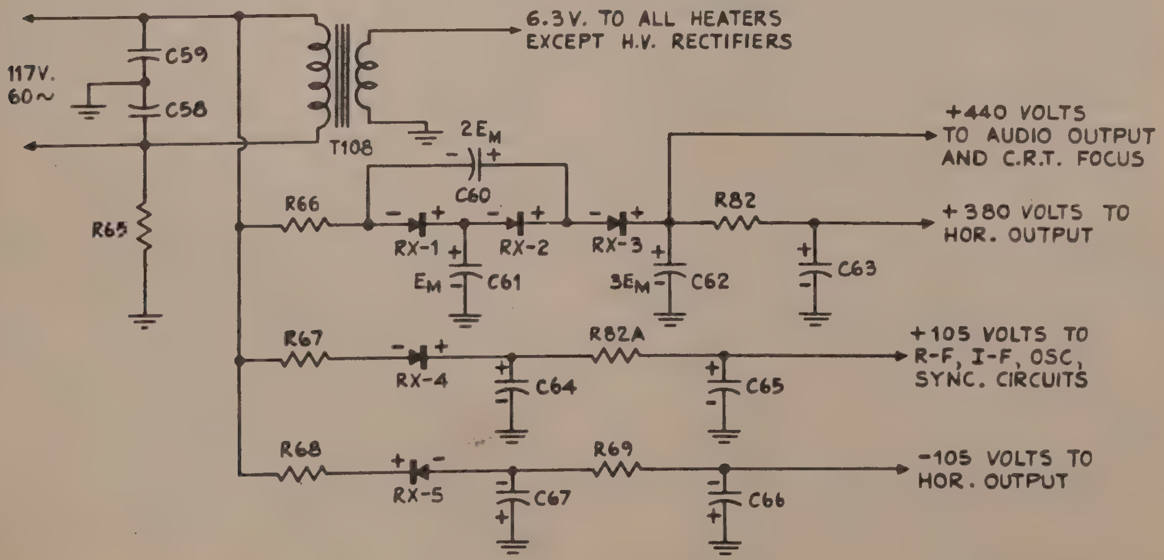

6.3V. TO ALL HEATERS EXCEPT H.V. RECTIFIERS
117V. 60~
C59
C58
T108
2E_M
C60
R65
R66
RX-1
RX-2
RX-3
E_M
C61
3E_M
C62
R82
C63
+440 VOLTS TO AUDIO OUTPUT AND C.R.T. FOCUS
+380 VOLTS TO HOR. OUTPUT
R67
RX-4
C64
R82A
C65
+105 VOLTS TO R-F, I-F, OSC, SYNC. CIRCUITS
R68
RX-5
C67
R69
C66
-105 VOLTS TO HOR. OUTPUT

FIG. 11-4.—The low-voltage power supply of the Consolidated Television model 2315. Note the use of the selenium rectifiers.

used in this supply is a Federal Type 403D2787, which is rated at 150 millamperes d-c output.

High-Voltage Power Supplies

There are three types of high-voltage power supplies that are used in television receivers. They are:

1. The 60-cycle high-voltage supply
2. The r-f high-voltage supply
3. The kick-back high-voltage supply

The basic circuit of a 60-cycle power supply is shown in Fig. 11-5. This circuit is very similar to the conventional circuits used in the low-voltage supply. The transformer T_1 steps up the 60-cycle line voltage to the high voltage required, usually in the order of several thousand volts. This voltage is rectified by the half-wave rectifier tube and then filtered by the *RC* filter. The current required by the anode is usually about 200 microamperes, therefore the voltage drop across the resistor is not too great. The resistor *R* should be quite large, at least a megohm, and the capacitors as small as will give satisfactory filtering. Large values of filter resistance rather than large values of capacitance are used to obtain the required filtering. The advantage of this is the reduced cost and size of the capacitors; in addition, the smaller the size of the filter capacitor, the less the danger of fatal shock if one accidently comes in contact with a charged or a partly discharged capacitor. *Nevertheless, the high-voltage filter capacitors should always be discharged before any measurements are made so as to guard against the possibility of an open bleeder circuit.* Above 400 volts, the transformer necessary for this type of power supply becomes heavy, bulky, and expensive. Also, the danger of fatal shock becomes greater. Because of these objections, this type of power supply is generally not used except in the smaller television receivers, usually those employing a 7-inch picture tube or smaller.

Most of the disadvantages of the 60-cycle supply are eliminated by the use of the r-f high-voltage power supply. A block diagram of this type of supply is shown in Fig. 11-6. Instead of using the 60 cycles from the power line for the input of the step-up transformer, a separate r-f oscillator is used to supply the input of the step-up transformer. The frequency of this oscillator is usually in the range of 50 to 500 kc. The transformer can be made much lighter and smaller than a 60-cycle power transformer because of the high frequency used. It is also much easier to filter this higher frequency, therefore, the filter capacitors can be made smaller. The oscillator is capable of delivering little more power than is actually required by the anode of the picture tube; this is usually about two watts. The regulation of the power supply is, therefore, very poor; that is, the output voltage falls off rapidly with an increase in load. This is a very desirable feature in a high-voltage power supply. If a serviceman accidentally came in contact with a lead carrying the high voltage, he would load down the power supply and the voltage would drop, lessening the chances of a fatal shock. However, this does not mean that it is safe to touch a high-voltage lead. Do not work on *any* high-voltage circuits unless the power has been turned off and the high-voltage filter capacitors have been discharged.

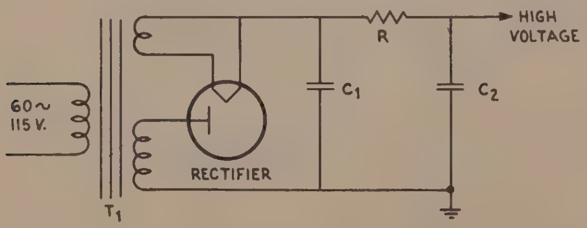


Fig. 11-5.—The basic 60-cycle power supply circuit. Here large values of filter resistance rather than large values of capacitance are used to obtain the necessary filtering.

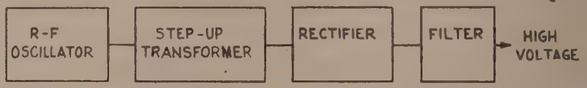


Fig. 11-6.—Block diagram of the r-f high-voltage supply, the frequency of the oscillator being in the range of 50 to 500 kc.

The r-f power supply used in the Belmont models 22A21 and 22AX21 is shown in Fig. 11-7. The oscillator circuit uses a 6V6 as a tuned plate, tickler feedback oscillator. The frequency of oscillation is determined by the setting of the trimmer capacitor *C*86. The secondary winding together with a winding for the filament of the rectifier tube are both wound on the same form as the oscillator primary and tickler coil, thus combining the oscillator coil and the step-up transformer. The secondary winding has many more turns than the primary winding, therefore the voltage induced in the secondary will be many times greater than the voltage applied across the primary by the oscillator. The frequency of the oscillator is determined by the setting of the trimmer capacitor *C*86. The resonant frequency of the secondary winding is determined by the shunting capacitance due to the rectifier tube. The voltage across the secondary will be greatest when the frequency of the oscillator is equal to the resonant frequency of the secondary winding in parallel with the shunting capacitance. The circuit is designed so that the maximum voltage that is obtainable at resonance will be greater than that required by the television picture tube. In actual operation, the trimmer is adjusted until the output voltage falls off to the required value. The *LC* combination of *L*6 and *C*84 makes up a decoupling filter to keep r.f. out of the low-voltage power supply.

The method of heating the high-voltage rectifier tube is interesting. A small winding on the same form as the other windings has sufficient r-f voltage induced in it to heat the filament of the rectifier tube. This has a decided advantage over heating the filament with 60-cycle current. If 60 cycles were used, the high-voltage filter would have to filter out the 60 cycles due to the filament current. This would mean very large capacitors. By heating the filament with r.f., the filter has to take care of only the high-frequency r-f voltage. This is very easily done, as can be seen from the small capacitors used in the filter circuit. In this circuit, *C*89 and *C*90 are only 0.001 μf each.

The third type of high-voltage power supplies used in television receivers is the kick-back type of supply. The distinguishing feature of this power supply is that the rectifier plate voltage is obtained from the horizontal deflection system during the retrace or flyback of the sweep, rather than from the a-c line as in the case of the 60-cycle supply or from a separate high-frequency oscillator as in the case of the r-f power supply. A block diagram of this type of supply is shown in Fig. 11-8. The high-voltage supply in the G.E. model 802 is typical of the kick-back high-voltage supplies used in television receivers.

To analyze the operation of this circuit, we must see how the horizontal output tube, a 6BG6G, and the damping tube, a 6AS7G, generate a pulse that excites the step-up transformer, *T*25, thus providing a high voltage which is rectified by the 8016, as shown in Fig. 11-8. The 6BG6G is a beam tetrode, capable of handling heavy current; its main function is to supply current with the proper waveform to the horizontal sweep coils so that a horizontal trace of the proper length is applied to the viewing tube. The 6AS7G is a dual triode, having an amplification factor of only 2.1 and a plate resistance of 140 ohms when the two sections are connected in parallel. This tube damps the high-frequency oscillation set up during the flyback period of the electron beam.

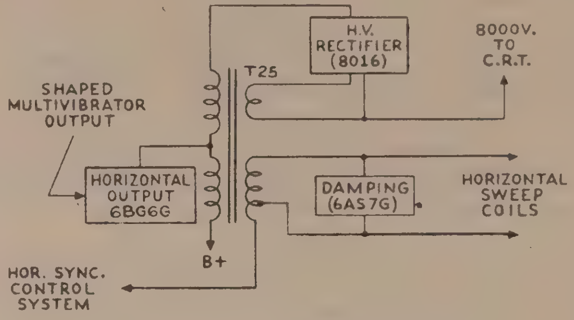


Fig. 11-8.—The "kick-back" type of power supply such as is used in the G.E. model 802. Here the rectifier plate voltage is obtained from the horizontal deflection system during the flyback periods.

The usual high-voltage rectifier, such as an 878, requires considerable heater power and is not designed for r-f operation. The 8016 diode was developed for this purpose and requires only a quarter of a watt for the heater. Fig. 11-8 shows that the heater voltage for the 8016 is obtained by a very small secondary winding on the transformer, *T*25. The 10BP4 picture tube is scanned by a magnetic coil system, but only the horizontal sweep coils enter our study of the high-voltage power supply.

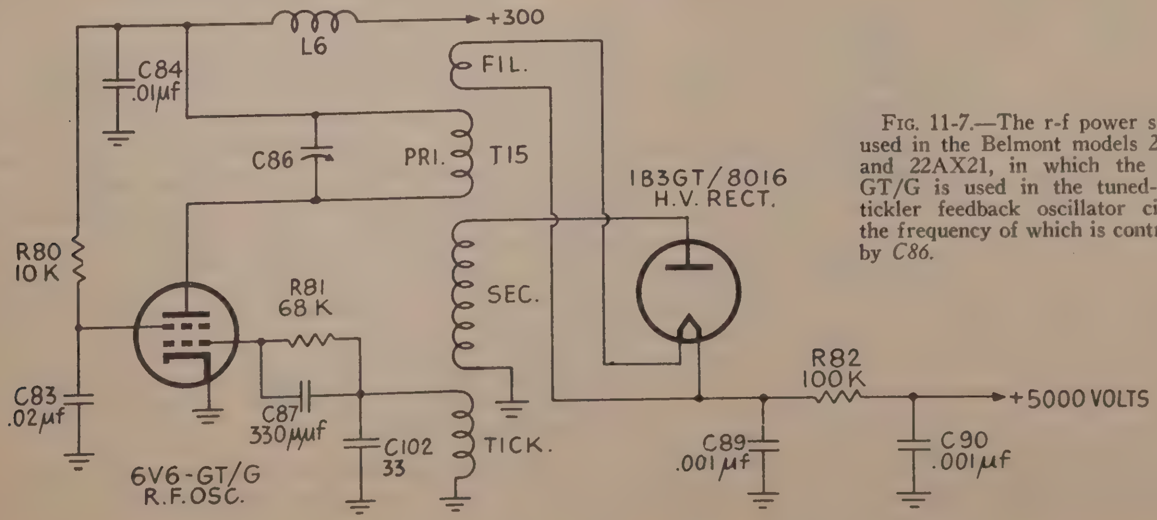


Fig. 11-7.—The r-f power supply used in the Belmont models 22A21 and 22AX21, in which the 6V6-GT/G is used in the tuned-plate, tickler feedback oscillator circuit, the frequency of which is controlled by *C86*.

For proper scanning, the current in the sweep coils must have a saw-tooth form, such as shown in Fig. 11-9, the forward trace across the screen occurring during period *a-b* and the retrace or flyback during the much shorter period *b-c*. The forward trace must be as linear as possible so that distortion is prevented.

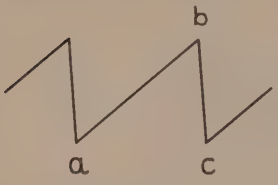


Fig. 11-9.—The current in the sweep coils must have a saw-tooth form, *a-b* being the trace period and *b-c* the retrace.

The sweep trace takes place when the output of the 6BG6G tube is applied to the horizontal sweep coils, and the transformer *T*25 is so designed that a proper impedance match exists between the tube and the coils. When the 6BG6G stops conducting, the components to the right of *T*25 (See Fig. 11-11) are excited into violent oscillation, and the oscillation is used to obtain the rapid flyback, *b-c* in Fig. 11-9. Any oscillation beyond the first half cycle is undesirable, as it will affect the linearity of trace *a-b* in Fig. 11-9, and the 6AS7G is used to damp it out. Very little of the magnetic energy is consumed by the flyback, and the collapsing field produces a positive voltage pulse on the primary of *T*25. This pulse is stepped up by the additional winding shown on the primary; it is then rectified by the 8016 and delivered to the viewing tube.

The detailed functioning of each component may now be examined, keeping the previous discussion in mind. The input to the 6BG6G is obtained from a multivibrator (6SN7-GT) which is a two-tube oscillator that normally produces a rectangular output. However, the network composed of the capacitors and resistors in the input circuit of the 6BG6G is so designed that we obtain a saw-tooth waveform with a negative pulse, such as shown in Fig. 11-10. This is the input that is applied to the grid of the 6BG6G. The collapsing field in the horizontal deflection coils produces a positive voltage pulse on the primary of *T*25 and consequently on the 6BG6G plate. The negative pulse on the 6BG6G grid, *a-b* in Fig. 11-10, ensures that the 6BG6G is cut off during the flyback period, in spite of the high plate voltage.

During the trace period (*c-a* in Fig. 11-10) when the 6BG6G is conducting, the saw-tooth grid voltage produces a saw-tooth plate current. The effective plate load consists of the horizontal deflection coils, *T*25 being used only as an impedance transformer to match the coils to the tube. Since the plate load is inductive, if a sudden change in current takes place, we obtain a high-voltage pulse. This sudden change takes place during *a-b* of Fig. 11-10. During *c-a* of Fig. 11-10 the energy supplied to the yoke builds up, and continues to do so until the negative pulse occurs. At this instant, when the

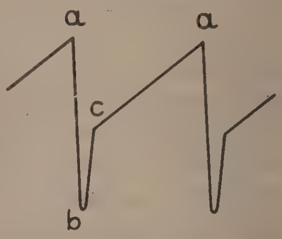


Fig. 11-10.—The saw-tooth waveform with a negative pulse, that is applied to the grid of the 6BG6G tube.

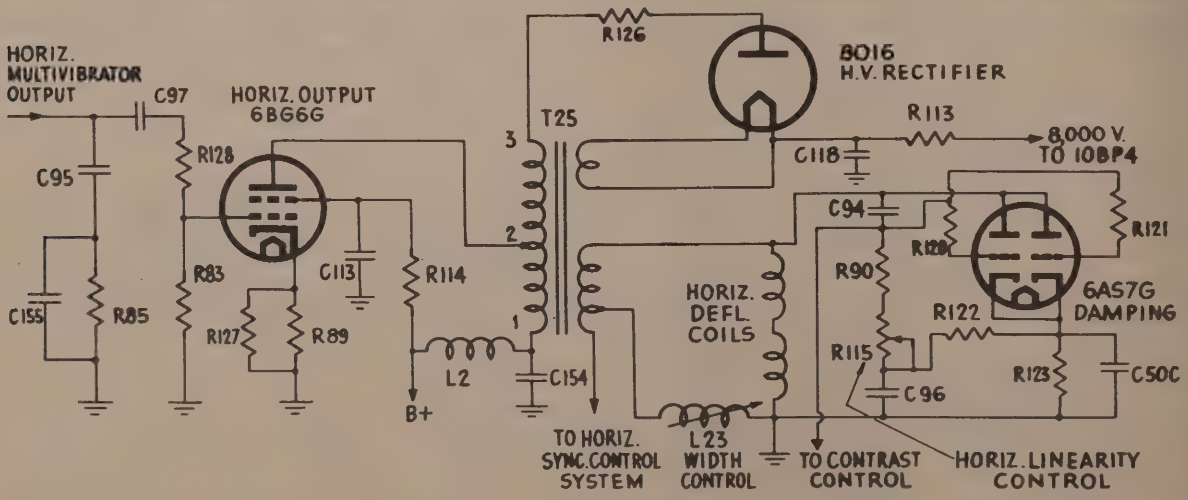

HORIZ.
MULTIVIBRATOR
OUTPUT
C97
HORIZ. OUTPUT
6BG6G
C95
R128
C155
R85
R83
C113
R114
R127
R89
L2
B+
C154
T25
3
2
1
R126
8016
H.V. RECTIFIER
R113
8,000 V.
TO 10BP4
C118
C94
R120
R121
R90
6AS7G
DAMPING
HORIZ.
DEFL.
COILS
R122
R115
R123
C50C
C96
TO HORIZ.
SYNC.CONTROL
SYSTEM
L23
WIDTH
CONTROL
TO CONTRAST
CONTROL
HORIZ. LINEARITY
CONTROL

Fig. 11-11.—Schematic diagram of the high-voltage supply circuit of the G.E. model 802. The block diagram of this circuit is shown in Fig. 11-8.

6BG6G is cut off, the components between *T*25 and the 6AS7G are shocked into violent oscillation (*L*23, horizontal deflection coils, part of *T*25, and distributed capacitances). During the first half cycle, which is negative, the current in the horizontal deflection coils reaches a maximum in the direction opposite to that in which it is flowing and the flyback consequently takes place. At the end of the first half cycle, the voltage starts to go positive. The 6AS7G now conducts, since its plate is positive, and the oscillation is rapidly damped.

Transformer *T*25 has four functions. Its main function is to transform the inductance of the horizontal deflection coils to a value that meets the operating conditions of the 6BG6G. A small secondary tap supplies heater voltage for the 8016. It also takes the inductive "kick" voltage from the collapsing magnetic field flyback and places it on the primary. Finally, this voltage pulse on the primary is raised by autotransformer action to 8000 volts and is applied to the plate of the 8016. (In an autotransformer the input is applied to a portion of the winding, terminals 1-2 of Fig. 11-11, and the output is taken across the entire winding, terminals 1-3. It has these advantages over the ordinary two-circuit transformer: better voltage regulation, greater efficiency, and smaller size; it does have disadvantages, such as lack of d-c separation, which prevent its more universal use.)

The 8016 rectifies this high-voltage pulse. Due to the high frequency at which this takes place, a 500 $\mu\mu f$ capacitor, *C*118, is sufficient for filtering; and as its small value means small energy storage, there is consequent reduction of danger when servicing the high-voltage supply. The 8000-volt output is applied to the picture tube through the filter section *C*118 and *R*113.

Some of the signal on the secondary of the transformer *T*25 is used to operate the horizontal sync control system. This system is also known as automatic frequency control (afc) and is discussed in detail in chapter 8. The negative bias developed in the grid circuit of the 6AS7G damping tube is used to supply a steady negative d-c voltage for the contrast control circuit. The contrast control varies the gain of the i-f amplifiers by varying the grid bias on the first two stages and thereby controls the contrast of the picture. This is discussed in chapter 6.

The GE models 901 and 910 are projection type television receivers. In order to obtain an image on the picture tube bright enough so that it can be projected onto a screen, very high voltages on the second anode are necessary. The 5TP4 projection tube used in this receiver requires 27,000 volts on the second anode and about 5000 volts on the first anode. It would be very difficult to obtain these very high voltages by the use of only one rectifier. Assuming that a kick-back type of supply is used, the high-voltage rectifier tube, the horizontal output transformer, and all associated wiring would have to withstand this very high voltage. This, of course, would increase the cost of the power supply tremendously. For these reasons the high-voltage supplies for the projection type picture tubes generally are of the voltage, multiplying type. The high-voltage supply of the GE model 901 consists of a kick-back type of power supply plus a voltage multiplier. Except for the voltage multiplier, this supply is very similar to the kick-back supply of the GE model 802 (see Fig. 11-11) which we have already analyzed. Therefore it will be sufficient for us to discuss the theory and analyze the operation of the voltage multiplier circuit, a simplified schematic of which is shown in Fig. 11-13.

First, let us look at the basic circuit shown in Fig. 11-12. In this circuit, we have a source of d-c voltage as represented by the battery and a switch which will put the capacitor C_1 either across the battery for charging or across *R* and C_2 for discharging. At the beginning of the experiment we will assume that neither capacitor has any charge on it. Now, we will place the switch in position 1 for a very short time. Since capacitor C_1 does not have any resistance in series with it, it will charge up almost immediately to the full voltage of the battery, which is *E*. Then the switch is turned to position 2 and held there for a much longer time than in position 1, since the resistor in series with the capacitor C_2 prevents it from charging up as fast as it would without the resistor. While the switch is held in position 2, some of the charge from C_1 will go to C_2. If the two capacitances are equal, the voltage across each will become half of *E*. Now, suppose the switch is again placed in position 1 for a very short time. Capacitor C_1 will again charge up to *E*. When the switch is placed in position 2, C_1 will discharge into C_2 until the voltage across each capacitor is equal to ¾ of the voltage *E*. If this is kept up for a number of times, the voltage across C_2 will become greater and greater until eventually both capacitors are charged up to the voltage *E*. Actually the voltage across C_2 never

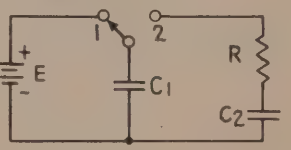


FIG. 11-12.—Equivalent circuit showing charging action occurring in the voltage multiplier shown in Fig. 11-13.

becomes exactly equal to *E,* but it is very close to it, so that for all practical purposes we may say it becomes equal to *E.*

If we replace the components of the circuit of Fig. 11-12 with the corresponding components of the simplified schematic shown in Fig. 11-13, we will have the first stage of the voltage multiplier used in this high-voltage supply. The transformer *T*202 and the first high-voltage rectifier, *V*207 are the equivalent of the d-c supply in the basic circuit of Fig. 11-12. The capacitor *C*214 corresponds to C_1, and the *RC* combination of *R*219-*R*222 and *C*211 corresponds to *R* and C_2 in Fig. 11-12. The switch action in the basic circuit is performed by the pulse that is developed in the primary of the transformer during flyback. The capacitance C_g represents the capacity between the second anode of the picture tube and the external conductive coating which is grounded. This capacity is usually around 250 μμ*f* and at the sweep frequency is large enough to be used as a filter capacitor.

When the flyback occurs, a positive pulse is produced in the primary of the transformer *T*202. Since the flyback period is a very small part of the entire sweep cycle, the pulse will be a comparatively short one, but due to the step-up in the transformer it will be about 8000 volts at the output of the transformer. Let us call the peak voltage of this pulse E_m. This voltage can be applied between the plate of the first rectifier tube, *V*207, and ground since the other end of the primary of the transformer *T*202 is connected to *B*+, which is an effective ground at the horizontal sweep frequency due to the large filter capacitors in the low-voltage supply. Since this pulse is positive, the rectifier *V*207 will conduct and charge *C*214 to the peak voltage E_m. During the forward trace, *C*214 will discharge through *R*219-*R*222 to charge *C*211 to E_m. The discharge path of *C*214 is through *R*219-*R*222, *C*211, *R*215, the primary of the transformer, *L*204, *B*+ and then to ground. When we say that *C*211 will charge up to E_m, we mean after a number of pulses in much the same manner as was done in the basic circuit in Fig. 11-12.

On the first positive pulse after the capacitors are fully charged, capacitor *C*211 will act in series with the pulse and a total voltage of $2E_m$ will be applied to the second rectifier tube, *V*208. This tube will conduct and will charge the series combination of *C*215 and *C*214 to a voltage of $2E_m$. Since the capacitances are equal, each one will be charged to a voltage of E_m. The polarity of these voltages is indicated in Fig. 11-13. During the forward trace, the capacitors *C*214 and *C*215 will discharge through the resistors *R*223-*R*226 to charge the capacitors *C*212 and *C*211. The rest of the discharge path is the same as in the first stage. *C*212 and *C*211 will now charge up to a total voltage of $2E_m$, but since they are equal, the voltage across each will be E_m.

On the next positive pulse, *C*211 and *C*212 will act in series with the positive pulse and apply a total voltage of $3E_m$ to the plate of the third rectifier tube *V*209. This tube will conduct and charge

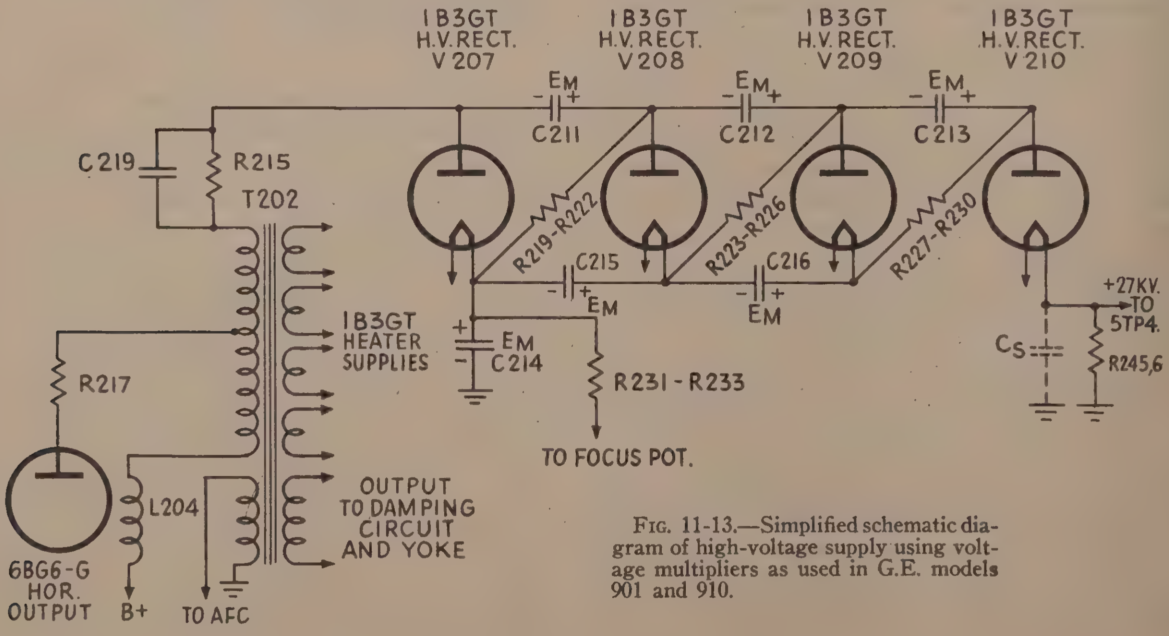


FIG. 11-13.—Simplified schematic diagram of high-voltage supply using voltage multipliers as used in G.E. models 901 and 910.

the capacitor $C216$ in a similar manner to the second stage. During the forward trace the three capacitors $C214$, $C215$, and $C216$ will discharge through $R227$-$R230$ to charge $C213$ to E_m. On the the next positive pulse $C211$, $C212$, and $C213$ will all act in series with the pulse so that a total voltage of $4E_m$ is applied to the plate of the fourth rectifier tube $V210$. This tube will conduct and produce an output voltage of nearly $4E_m$.across the bleeder resistors $R245$,6 and the shunting capacitance C_8. The output voltage will not be equal to quite $4E_m$ due to the current drawn by the picture tube and the voltage drops in the rectifier tubes. The output voltage is usually about $3.7E_m$ instead of the theoretical $4E_m$. The first anode of the 5TP4 requires a voltage of only about 5000 volts as compared with the 27,000 volts required by the second anode. The voltage at the cathode of the first rectifier tube $V207$, which is about 7000 volts, is tapped off through the resistors $R231$-$R233$ and is applied to the focus control potentiometer which supplies the voltage for the first anode.

CHAPTER 12

ALIGNMENT AND SERVICING

By William Bouie and Henry Chanes

When defects occur in a television receiver caused by misalignment or breakdown of parts, the defect must be isolated and corrections made to restore normal operation of the set in the least amount of time. This chapter is written with that end in view.

With experience, the average radio serviceman usually works out his own pet troubleshooting procedure with which he can quickly find the defects in inoperative radio receivers. This also applies to troubleshooting television receivers, and it is hoped that the information contained here will be of considerable help in establishing a fast working method consistent with the circuit specifications of the receiver being serviced. As regards alignment, each television manufacturer provides alignment data for his particular receiver and it is not intended that the general information contained herein should be substituted for these specific alignment data. However, if for any reason the manufacturer's alignment procedure is not available, this general procedure outlined here will enable the serviceman to align the set.

The troubleshooting section covers localizing methods used to isolate the defect to a particular section and stage. For television receivers incorporating conventional circuits, best results will be obtained by following the localizing procedure presented here. However, occasions may arise where an alternative method may prove more satisfactory. A troubleshooting chart is included at the end of the chapter to facilitate further investigation into the causes of probable breakdowns.

ALIGNMENT

A television receiver is able to select one particular station at a time to the exclusion of all other stations operating on different channels in a manner similar to the conventional superheterodyne radio receiver. Accurate tuning to a particular channel and the reception of high-quality sound and picture detail is dependent upon optimum adjustment, or alignment, of the various tuned circuits within the television set. These tuned circuits, which consist of acceptance and rejectance circuits (the latter known as traps), are incorporated into the r-f, mixer-oscillator, sound, and video sections of the receiver. An examination of the schematic and parts-layout diagrams for the particular receiver is required to locate the various trimmers and slugs to be adjusted.

A simple block diagram of a conventional television receiver is shown in Fig. 12-1. This diagram shows the television receiver to consist of a number of individual sections performing specific functions. For the television receiver to operate properly and with the greatest efficiency, the tuned circuits in each section must be adjusted to certain frequencies. Adjustment of these tuned circuits in a definite sequence constitutes what is known as alignment.

From the block diagram, it can be seen that the function of the r-f section of the receiver is to select the desired station and to amplify the incoming signal before passing it on to the mixer stage. In order to perform both of these functions, it is essential that the tuned circuits present in the r-f unit be tuned accurately to the frequency of the incoming signal. The oscillator coil for each channel must be adjusted to produce a frequency which, upon being beat with the frequencies of the picture and sound carriers of the selected television channel, will produce the i.f.'s for the sound and video sections of the receiver.

Correspondingly, the sound section and the video section of the set require adjustment in order to pass their respective i.f.'s through to the detector stages. Additional tuned circuits are placed in the video section to keep interfering frequencies from appearing in the picture with resulting distortion effects. These latter circuits are known as sound or video traps, depending upon the particular interfering frequencies to be suppressed.

It must be determined whether the faulty operation of a television receiver is due to poor alignment or to some defect in the set before any alignment is attempted. In entirely too many instances there is a tendency to blame poor receiver operation upon the alignment and to upset a perfectly good alignment without first having investigated for defects in the set. This tendency should be discouraged and an examination of the set should be made to determine the true cause of the defective operation before any alignment is attempted.

An incorrect alignment condition in a television set is generally accompanied by low sensitivity, loss of sound or picture or both, and distortion in sound and/or picture. Misalignment of the trap circuits is indicated by interference in the picture caused by sound frequencies of the channel being received, or by picture and/or sound frequencies of an adjacent channel. These conditions may occur on one or more channels and may be present in varying degrees depending upon which tuned circuits are out of alignment and the extent of the misalignment.

Considerable time may be saved by analyzing observed indications to determine which sections of the receiver are out of alignment. For example, if misalignment is indicated on only one channel, it would be safe to assume that only the local oscillator or the r-f circuit of this channel requires alignment. Proper response on the other channels shows that the sound and video sections are properly aligned and do not require realignment. Conversely, if the receiver shows low sensitivity to sound and/or picture on all channels, misalignment of the sound or video i-f sections is indicated because misalignment of this part of the receiver will drop the sensitivity on all channels uniformly. While it is possible for the alignment on all the channels to be out, the more probable condition is that the i-f sections require realignment, and these sections should be checked first.

There are a number of different factors which make realignment necessary at more or less frequent intervals. Perhaps the factor which is responsible for more realignment jobs than any other is the change in the characteristics of the components associated with the tuned circuits of the receiver. Due to vibration, the movement of parts, and the effects of humidity, temperature, and age, the capacitance and inductance associated with these tuned circuits change their values, and the tuned circuits go out of alignment. Aside from changes in the tuned circuit itself, there are a number of other factors which make realignment necessary. Among these can be mentioned the movement of r-f and i-f wiring, since the movement of these leads changes the relative capacitances and inductances associated with the tuned circuits. A slight change in position of the wiring may cause a channel, or even the entire receiver, to be inoperative since the leads constitute a large part of the inductance and capacitance of the tuned circuits.

Particular mention in this connection must be made of the importance of using exact replacement parts where replacements of resistors, capacitors, and other parts becomes necessary in the r-f amplifier, mixer, oscillator, or i-f sections of the television receiver. The use of a part which has the same rating as the part being replaced but which has different physical characteristics or size will sometimes throw the receiver out of alignment, and in other cases even cause instability and oscillation. It also should be kept in mind that many replacements parts are rated at low frequencies where stray capacitances and inductances are not significant. If an exact replacement part is not used, these characteristics must be taken into consideration. In addition, replacement of tubes in the front end and i-f sections of some receivers may cause sufficient change in the tuned circuit to require realignment. Occasionally when replacing an oscillator tube, it may be necessary to try several tubes before one is found that will operate properly in the oscillatory circuit.

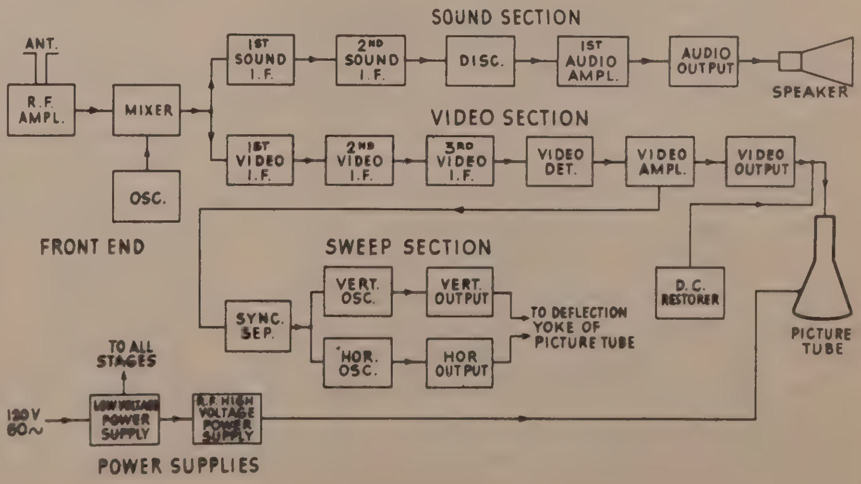


FIG. 12-1.—Block diagram of a television receiver, which can be divided into five sections: the radio frequency or front end, the sound, video, sweep and sync sections, and the power supplies. In order for the set to function properly, the tuned circuits in these sections must be aligned to certain frequencies.

Equipment Required

The basic test equipment required for the alignment of a television receiver should consist of the following:

1. Sweep Generator.
2. Marker Generator.
3. Oscilloscope.
4. Vacuum-tube Voltmeter (VTVM)

The sweep generator is an r-f signal generator whose output frequency can be made to vary continuously over a range of frequencies. This variation in frequency over a specific band is automatic and is called "sweeping." This action, in effect, constitutes f.m., and in some manufacturers' literature the sweep generator is referred to as an f-m signal generator. This sweep generator in conjunction with an oscilloscope facilitates alignment by providing a visual indication of the response characteristic of the circuit being aligned.

The sweep generator should have the following characteristics:

a. The frequency range should cover all the television channels from about 40 megacycles to 220 megacycles. It must also cover the i.f.'s of the video and sound sections, which range from approximately 20 to 30 megacycles for later models of receivers. In early models, i.f.'s as low as 8 mc may be encountered. Accuracy of frequency calibration is not too important in this generator as the exact frequency will be determined by the marker generator.

b. The sweep width should be variable, with a maximum width of at least 10 mc. The output of the sweep generator should be constant over the entire range of frequencies being swept. If this is not so, it will be impossible to obtain a true response pattern.

c. The sweep generator should have an adjustable attenuator control in the output circuit with a maximum available output of at least 0.1 volt. Provision should be made for either a balanced or unbalanced output. A balanced output is desirable when coupling the generator to the antenna terminals of a receiver having a balanced input circuit. An unbalanced output is generally used when the generator is connected to other points in the television receiver.

d. Provision must be made in the sweep generator to supply a horizontal sweep voltage or sync signal for the oscilloscope. If a horizontal sweep voltage is provided, this voltage must be of the same frequency and waveform as the voltage used to modulate the carrier frequency of the sweep generator. Usually a phase adjustment control is included in this circuit to shift the phase of the sweep voltage supplied to the oscilloscope. The double trace on the oscilloscope, resulting from phase shift, can thus be resolved into a single response pattern.

The marker generator is an r-f signal generator used to supply marker pips on the response pattern being observed, in order to indicate the exact frequency of any point on the pattern. This generator also may be used to peak the i-f stages and align traps. It should contain the following features:

a. The marker generator should cover the same frequency ranges as the sweep generator, but in this case the accuracy of frequency calibration is of extreme importance. Frequency stability and calibration in the marker generator must be exceptionally good for proper alignment of television receivers.

b. An adjustable attenuator control should be incorporated in the output circuit enabling the output to be varied from 0 to 0.1 volt. In some generators the attenuator control is calibrated in microvolts; in others the level of the output is indicated by a meter.

c. In addition to supplying an unmodulated signal, the marker generator must have provision for an a-m r-f output signal. The modulating frequency is not critical but most manufacturers provide a 400-cycle signal. A calibrated indication of the percentage of modulation up to approximately 80 percent is a desirable feature.

The oscilloscope used in the alignment of television receivers can be any of the conventional types used for test purposes. However, oscilloscopes with 5-inch tubes may facilitate the alignment by virtue of the larger pattern obtained.

The VTVM used in the alignment should have a low d-c scale of the order of three volts and an input impedance of about ten megohms. A zero-center scale will be found useful in balancing the discriminator output and in adjusting the local oscillator.

Alignment Procedures

In the actual alignment of a television receiver, the following pertinent facts should be borne in mind: Quite often response curves are obtained that appear to be correct but that do not indicate the true response of the circuit. If a marked change in the shape of the response curve is obtained with changes in output level or sweep width of the sweep generator, the curve is, in all probability, not a true indication of circuit conditions. This false response may be caused by oscillation or regeneration in the circuit, or by overloading of one or more of the i-f amplifiers. The solution for overloading is self-evident; always use the smallest signal possible to provide a suitable indication on the output device which should be used at maximum sensitivity, whether it be an oscilloscope or VTVM.

Regeneration and oscillation can be caused by too close coupling of the generator leads since pickup or radiation from these leads can cause undesirable interaction between the stage to which the generator is coupled and the other stages in the circuit. Regeneration or oscillation can be prevented by the use of well-shielded leads from the signal generator and by means of loose coupling. Sufficient coupling between the signal generator and the circuit may very often be obtained by merely clipping the generator lead to the insulated wiring of the circuit.

All that has been said with regard to the sweep generator is applicable to the marker generator as well. In addition, using the two signal generators together may increase the danger of regeneration and oscillation. This can be prevented by isolating the two signal generators from each other by keeping their connecting points at least one stage apart whenever possible. It is usually possible to leave the marker generator connected to the mixer grid while the sweep generator is moved progressively through the i-f section of the receiver. The output of the marker generator should be kept as low as possible; only a small marker pip is required on the oscilloscope. If the marker pip is too large, the response pattern will be distorted.

The response curve on the oscilloscope may appear in reverse, that is, the horizontal sweep line may be deflected either upward or downward depending upon the arrangement of the video detector circuit. The polarity of the pattern is of no consequence; either pattern is equally useful. The low-frequency end of the response curve may be on either the right or left side of the pattern. This is dependent upon the polarity of the horizontal sweep voltage supplied to the oscilloscope by the sweep generator. The low- and high-frequency extremes of the pattern should be determined by means of the marker generator and should be kept in mind when comparing the indicated pattern on the oscilloscope with reference patterns.

It is important that all of the test equipment and the receiver being aligned be bonded together and connected to a common ground.

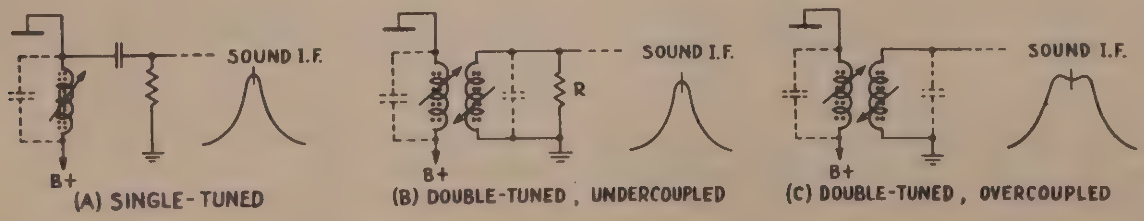
For over-all alignment of a television receiver, a definite sequence of alignment must be followed for several reasons. First, the response of some circuits depends upon the proper adjustment of others. Second, when an orderly sequence is followed, a complete section of the receiver can be aligned without too many changes in the settings of the test equipment used. This saves time and, in addition, will reduce errors that might occur when attempting to reset the signal generator exactly to a frequency previously used. This last reason is especially important when making adjustments in the trap circuits.

When manufacturers' alignment instructions are supplied, these instructions follow the sequence recommended by that particular manufacturer. It is our suggestion that this sequence should be followed. However, in the absence of manufacturers' data, the alignment procedure outlined here will be of considerable value. This procedure is arranged in the following sequence:

1. Sound i-f section.
2. Trap circuits.
3. Video i-f section.
4. Oscillator.
5. R-f amplifier and mixer.

Sound I-F Section

The coupling circuits in the sound i-f section may be either single- or double-tuned as shown in Fig. 12-2. The coils are usually slug tuned and are resonated with a shunt capacitance, which may be an actual capacitor or may be the output or input capacitance of the tube. The single-tuned circuit shown in (A) of Fig. 12-2 has a single-peaked response characteristic as shown. This response curve is broadened by the comparatively small value of resistance, *R*,



(A) SINGLE-TUNED (B) DOUBLE-TUNED, UNDERCOUPLED (C) DOUBLE-TUNED, OVERCOUPLED

FIG. 12-2.—The coupling circuits in the sound i-f section may be single- or double-tuned as indicated above. Note the similarity between the response curves of (A) and (B) and how the overcoupling of the double-tuned circuit in (C) results in a widened response curve.

which is in the grid circuit of the following stage. The circuit shown in (B) of Fig. 12-2 is an undercoupled, double-tuned circuit having the same essential characteristics as that shown in (A). As will be noted, the response curve is also similar. In (C) the circuit is also double tuned, but in this case it is overcoupled to provide a wide response characteristic. This curve has two peaks, with the sound i-f center frequency occurring between the two peaks.

Inasmuch as the circuits shown in (A) and (B) of Fig. 12-2 have a single peak, alignment can be performed with the marker generator and the VTVM. However, due to the double-peak response of the circuit shown in (C), these instruments alone are not sufficient to adjust this latter circuit, visual alignment by means of a sweep generator and an oscilloscope being required.

We will now consider the alignment procedure of a sound section having single-tuned coupling circuits. This procedure will also apply to the double-tuned, undercoupled circuit in (B), with the exception that two adjustments are made instead of one. The procedure is as follows:

1. Connect the marker generator to the grid of the mixer tube. Set the marker generator exactly to the sound i.f.

2. Connect the VTVM to the discriminator output circuit. Set the VTVM to a low scale.

> Caution: Adjust the signal output from the marker generator to a value sufficiently low to provide the minimum indication required on the VTVM. As the stages are peaked, the output of the marker generator must be reduced further. Do not advance the VTVM to a higher scale to prevent the meter pointer from going off scale due to alignment of the circuits. Always reduce the output of the marker generator, working with the least possible signal to insure correct results.

3. First detune the discriminator secondary and then adjust the discriminator transformer primary for maximum indication on the VTVM, then progressively adjust the i-f coupling units, working back from the discriminator stage to the first sound i-f stage. This procedure may be repeated for greater accuracy.

4. Adjust the discriminator transformer secondary for a balanced condition as indicated by zero reading on the VTVM. This zero point will lie between the positive and negative peaks and is quite critical. When the discriminator is properly balanced, positive and negative peaks will be obtained as the marker generator is tuned to either side of the sound i.f. These peaks should be of equal amplitude. This completes the adjustment of the sound i-f section.

At this point it is advantageous to check visually the response characteristic of the sound section by means of a sweep generator and the oscilloscope. To do this, connect the sweep generator to the grid of the mixer tube. The marker generator is left at the same point, but is now loosely coupled. This loose coupling may be accomplished by means of a small series capacitor or by merely clipping the marker generator output lead to the insulated grid lead.

Note: The two generators should never be connected directly together. Connect the vertical input of the oscilloscope to the discriminator output. The horizontal input of the oscilloscope is connected to the horizontal sweep voltage supplied by the sweep generator.

Set the sweep generator to the frequency of the sound i-f section with a sweep frequency of approximately two megacycles. The pattern shown in Fig. 12-3 should be obtained on the oscilloscope. The frequencies corresponding to points *A*, *B*, and *C* in the figure can be determined by marker pips from the marker generator. To obtain these marker pips, set the marker generator to the i.f. of the sound section and vary the marker frequency along the slope of the response curve from peak to peak to determine the frequency limits of the curve. These frequency limits are usually of the order of 200 to 600 kc apart, depending upon the particular receiver. If the response curve is not symmetrical or is of insufficient width, it will be necessary to retouch the discriminator adjustments.

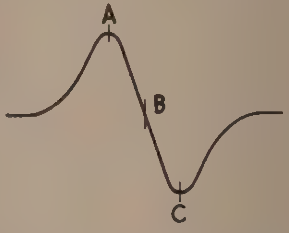


FIG. 12-3.—The response curve of the sound section obtained at the output of the discriminator. *A*, *B*, and *C* are pips obtained from the marker generator.

The alignment of the double-tuned, overcoupled circuit shown in (C) of Fig. 12-2 is as follows:

1. Connect the sweep generator to the grid of the mixer tube. Loosely couple the marker generator to the same point.

2. Set the two generators to the sound i.f. The sweep generator is set to a sweep of about two megacycles.

3. Connect the oscilloscope to the grid of the limiter tube. If no limiter tube is employed, completely detune the discriminator transformer secondary and place the oscilloscope across the discriminator output.

4. Adjust the primary and secondary of each coupling circuit in turn, working back from the last i-f tube to the first i-f stage. Proper response is indicated by a symmetrical curve peaked on either side of the sound i.f., as shown in (C) of Fig. 12-2. The object is to obtain maximum peaking of the curve while maintaining symmetry on either side of the sound i.f.

5. The oscilloscope is, if necessary, moved to the output of the discriminator and the discriminator transformer is adjusted for the response curve shown in Fig. 12-3. The primary of the discriminator transformer essentially determines the positions of points *A* and *C* on the response characteristic. The position of point *B*, which should correspond exactly to the sound i.f. is determined by the adjustment of the discriminator transformer secondary. Both of these adjustments should be retouched to obtain fine over-all response.

Trap Circuits

Trap circuits are used in the video section of the television receiver to reject frequencies that might cause interference in the picture. Fig. 12-4 shows a typical television channel with the two adjacent channels. Interference with the picture being received may be caused by its own sound carrier, the sound carrier of the next lower channel, or the picture carrier of the next higher channel. These carriers beat with the local oscillator and form i.f.'s which may lie within the pass band of the video i-f section of the receiver and, thereby, cause interference unless rejected by trap circuits. This is illustrated in Fig. 12-5. These interfering frequencies are known as accompanying sound, adjacent sound, and adjacent video frequencies. Most receivers utilize a trap in the video i-f section to eliminate adjacent and accompanying sound i.f.'s. Some receivers, in addition to these traps, employ a trap for the adjacent picture i.f.'s.

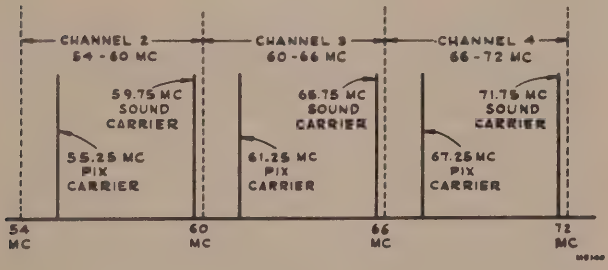


FIG. 12-4.—A typical television channel, No. 3, with adjacent channels 2 and 4.

Another form of interference may be caused by the beating together of the sound and picture i.f.'s of the channel being received. This beating occurs in the video detector and produces a difference frequency equal to 4.5 mc. In receivers having a wide response in the video amplifier stages, it may be necessary to employ traps to eliminate this interference. It cannot be too greatly stressed that the accurate setting of these traps is extremely important for interference-free operation of the television receiver. To insure the accurate setting of these traps, the signal generator must be set exactly to the rejection frequency of the trap being adjusted.

The following is a suggested procedure for aligning the trap circuits:

1. Connect the marker generator to the grid of the mixer tube.

2. Connect the oscilloscope to the output of the video detector.

3. Set the marker generator to the sound i.f. with modulation on. If the trap circuit is not properly adjusted, a signal will appear on the oscilloscope; this signal will be the modulating signal from the marker generator which has been detected in the video detector stage.

4. Adjust the accompanying sound trap for minimum signal indication on the oscilloscope.

5. Set the marker generator to the adjacent sound i.f. This frequency is 6 mc above the sound i.f. of the channel being received.

6. Adjust the adjacent sound trap for minimum output indication on the oscilloscope.

7. If the receiver has an adjacent video trap, set the marker generator to the adjacent video i.f. which is 6 mc below the picture i.f. of the channel being received. Adjust this trap for minimum output on the oscilloscope.

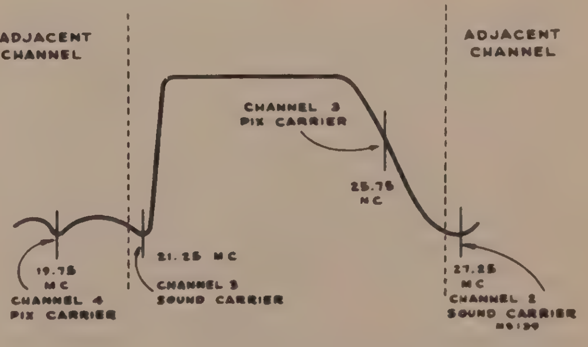


FIG. 12-5.—An over-all video response characteristic for channel 3, showing the interfering frequencies which must be eliminated by means of traps.

In elimination of the 4.5-mc video interfering frequency, connection of the marker generator and the oscilloscope will be determined by the location of the video trap in the receiver circuit. Usually the marker generator can be connected to the grid of the first video amplifier, with the oscilloscope being connected to the grid of the picture tube. With the marker generator set at 4.5 mc and modulated, a signal will appear on the oscilloscope if the trap is misaligned. If no indication is received on the oscilloscope, it may also mean that detection of the modulated signal is not occurring. In this case, use a crystal detector probe in series with the input lead of the oscilloscope in order to detect the modulation on the 4.5-mc signal input. With a signal indication on the oscilloscope, adjust the 4.5-mc video trap or traps for minimum indication.

Video I-F Section

The alignment method used in making adjustments in the video i-f section of the television receiver depends upon the type of interstage coupling used. There are two principal types of coupling circuits in use at the present time, the stagger-tuned and the overcoupled circuits. In most stagger-tuned circuits, each of the interstage coupling circuits is tuned to a different frequency. The over-all response characteristic of the i-f section is determined by the combination of the individual characteristics of each stage and will, therefore, have the required frequency bandwidth for the video i-f section. In most overcoupled circuits, the individual stages are all tuned to the same frequency. However, the coupling between primary and secondary of each interstage i-f transformer is great enough to produce a broad double-peaked response curve. This response curve can be further broadened by shunting the transformer windings with loading resistors to provide the required response characteristic.

In comparing the two types of coupling circuits, it should be noted that the individual stages in the stagger-tuned method of coupling are single peaked, while those of the overcoupled systems are double peaked. As in the alignment of the sound i-f section, the stagger-tuned coupling circuits, being single-peaked, can be aligned by means of the marker generator and the VTVM. Since the overcoupled circuits are double-peaked, the use of the sweep generator and the oscilloscope is required in order to obtain a visual response characteristic. A typical over-all video i-f response curve is shown in Fig. 12-5.

The following procedure is a suggested method for aligning a video i-f section containing stagger-tuned coupling circuits.

1. Connect the marker generator to the grid of the mixer tube.
2. Connect the VTVM to the output of the video detector; set to a low d-c scale.
3. Set the marker generator to the exact frequency of each i-f coupling circuit and peak each in turn, working progressively back from the video detector to the first i-f stage. The exact frequency of each coupling circuit will be determined by the specifications of the particular receiver.

NOTE: In order to check the over-all i-f response, it will be necessary to obtain a visual indication of circuit conditions following the above adjustment. This is done as follows:

4. Connect the sweep generator to the grid of the mixer tube. The marker generator is left at this point also, but is loosely coupled to prevent interaction.
5. Replace the VTVM with the oscilloscope at the output of the video detector.
6. Set the sweep generator to the center frequency of the i-f pass band with a 10-mc sweep. The over-all i-f response curve will appear on the oscilloscope. Use the marker generator to determine the frequency limits and identifying points on the pattern.
7. If necessary, retouch the alignment adjustments to obtain the proper over-all i-f response characteristic for the particular receiver.

In the alignment of a video i-f section incorporating overcoupled interstage transformers, a set of response characteristic curves will be required. The following procedure may be used:

1. Connect the marker generator to the grid of the mixer tube where it remains during the entire alignment procedure. The marker generator is used to provide marker pips for frequency identification.
2. Connect the oscilloscope to the output of the video detector.
3. Connect the sweep generator to the grid of the final i-f amplifier. Set the sweep generator to the center frequency of the i-f pass band with a 10-mc sweep.
4. Adjust the primary and secondary of the final i-f coupling transformer for the proper response pattern. Use the marker generator to identify limits and peaks of the response curve.
5. Move the sweep generator progressively back from stage to stage, adjusting each interstage transformer in turn until all are aligned for proper response.
6. In order to obtain the over-all i-f response, the sweep generator is connected to the mixer grid. The i-f response characteristic obtained should conform to the reference characteristic. Note particularly the position of the video i-f carrier on the response curve as indicated by the marker pip. The video i-f carrier

should occur about halfway up the sloping side of the response curve. Also check to see that the trap frequencies are outside of the pass band and properly attenuated. If these frequencies are not attenuated, realignment of the trap circuits will be required.

Local Oscillator

The local oscillator used in television receivers is operated at a frequency that is higher than the carrier frequencies of the television channel. The signal of the oscillator beats with the incoming sound and video carriers to produce difference frequencies which are equal to the sound and video i.f.'s of the receiver.

For example, in a typical television receiver, the video i.f. may be equal to 26.6 mc and the sound i.f. equal to 22.1 mc. When this receiver is tuned to channel two, the local oscillator will operate at a frequency of 81.85 mc, since the video carrier is 55.25 mc and the sound carrier is 59.75 mc.

Since the receiver i-f sections have already been aligned for proper response, the frequency of the local oscillator must now be adjusted so that the proper i.f.'s can be produced. This adjustment is especially critical for the sound section, as the sound i-f section has a limited bandwidth in the order of several hundred kilocycles as compared to the 3- or 4-megacycle bandwidth of the video i-f section. An error of a few hundred kilocycles in the frequency of the local oscillator will have a relatively small effect on the video response but it may be sufficient to completely eliminate the sound. Because of this, the sound i.f. generally is used as a reference when adjusting the oscillator. The sound i.f. is also used as a reference because the discriminator provides an accurate indication of frequency variation. It should be understood that if the receiver has been properly aligned up to this point, adjusting the oscillator for the correct sound i-f response will automatically provide the correct video i-f response because of the fixed 4.5-mc difference between the sound and video carriers. The following procedure can be used for aligning the oscillator:

1. Connect the marker generator to the antenna terminals of the receiver. If the receiver has a balanced input circuit, the output of the marker generator should also be balanced.

2. Connect the VTVM to the output of the discriminator.

3. Set the marker generator to the exact sound carrier frequency of the channel being aligned. The accuracy of this frequency setting is very important.

4. Adjust the oscillator frequency for zero output from the discriminator. This zero will lie between a positive and a negative peak. The adjustment is critical and must be carefully performed.

5. Repeat steps 3 and 4 for each channel to be aligned.

If a sufficiently accurate signal generator is not available, the oscillator may be aligned by using the signal from the television station. To do this, substitute an antenna for the signal generator and set the receiver to the channel to be aligned. If the station is on the air at the time, perform steps 2 and 4 in the above procedure.

R-F Amplifier and Mixer

The number of tuned circuits and adjustments required in the r-f amplifier and mixer circuits depends upon the particular receiver design. A number of receivers employ an untuned antenna input circuit, but have a double-tuned coupling circuit between the r-f amplifier and the mixer. In other receivers both coupling circuits are tuned, but only single tuning in each circuit is used. These circuits can be tuned to the different channels by varying either the circuit capacitance or inductance, or by switching in different values of inductance or capacitance. The circuits are tuned to provide maximum response to each channel, consistent with a sufficiently wide bandwidth to pass all frequencies of the television channel. Some receivers require adjustments for each channel, while others may require adjustments for only one or two channels; the other channels being preset in accordance with the design of the receiver. While the exact aligning method of the tuned circuits in the r-f amplifier and mixer circuits may vary somewhat with different receivers, the following general procedure can be used with satisfactory results.

1. Connect the sweep generator to the antenna terminals. If the receiver input is balanced, use the balanced output from the sweep generator. Set the receiver to the channel to be aligned.

2. Loosely couple the marker generator to the antenna terminals. Do not directly connect the two generators. The marker generator is used to provide marker pips for frequency identification.

3. Connect the oscilloscope to the output of the video detector.

4. Set the sweep generator to the center frequency of the channel with a 10-mc sweep. The over-all video response characteristic should appear on the oscilloscope.

5. Adjust the tuned circuits, working back from the mixer to the antenna terminals, for maximum indication consistent with proper response. The re-

sponse pattern obtained should coincide with the over-all video i-f response characteristic previously obtained. A typical example of this response curve is shown in Fig. 12-5.

In some receivers, it is possible to align the tuned circuits of the r-f amplifier and mixer by observing the response pattern of these tuned circuits alone, rather than the over-all i-f response curve. With this method, the oscilloscope is connected to the mixer tube instead of to the video detector. Detection occurring in the grid circuit of the mixer tube will provide a signal for the oscilloscope. The sweep and marker generators are used as before, but the response pattern now obtained should be wide enough to include both picture and sound frequencies of the channel. The adjustments are made for maximum amplitude of the response curve consistent with sufficient bandwidth, as shown in Fig. 12-6.

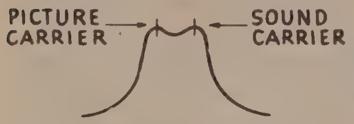


FIG. 12-6.—The response pattern of the tuned circuits of the r-f and mixer circuits.

TROUBLESHOOTING

While troubleshooting and repair of a television receiver may seem at first to present some rather insurmountable problems to the average serviceman, in actuality he will find himself dealing with many circuits which are common to conventional radio receivers. With a good basic knowledge of radio receiver circuits, and an understanding of the circuits peculiar to television receivers, which have been covered in this book, the servicing of these receivers should present no extraordinary problems. In the final analysis, a television receiver can be reduced basically to the usual electronic circuit components — tubes, coils, resistors, and capacitors — with which every serviceman is familiar.

Of course, the operating frequencies used in television are much higher than those used in radio receivers, necessitating greater care in checking circuits and replacement of parts. This is especially true of the circuits in the front end of the receiver which operate at the high frequencies of the various television channels. Also, the higher operating voltages used in some circuits which range from 7,000 to 30,000 volts, demand more than usual caution when checking or servicing these circuits.

Safety Precautions

In view of what has just been said about cautions in regard to high voltage, a few additional precautions are in order at this point.

1. Do not attempt to remove the chassis from its cabinet without *first turning off the power switch and removing the power plug from the a-c outlet.*

2. *Handle the picture tube with extreme care.* When placing it on the work bench, make sure that it is placed so that it cannot roll off the bench or bump against other objects. If the supporting mount is not removed from the cabinet with the tube, a jig should be fashioned to support the tube on the work bench. Use a thick soft padding under the tube when standing the tube on its face, to prevent scratching of the surface. Retain the cartons from new picture tubes to protect any tubes not in immediate use.

3. Before doing any soldering within the set, *be sure the power is turned off and the plug removed from the a-c outlet.* Discharge *all filter capacitors* by shorting them to ground by means of an insulated lead. Always connect the lead to ground first.

4. As a further caution in regard to the picture tube, *do not touch the high-voltage terminal on the tube even though the tube is disconnected.* Some tubes have a coating on the outer as well as on the inner surface which is usually the second anode. These two coatings act as the two plates of a capacitor and will develop a charge that may last for a considerable time if it is not discharged.

5. Do not attempt to reach into tight places inside the television cabinet to make any adjustments. In other words, *always allow sufficient room in which to work.* Also, *do not stand* on chairs, tables, or insecure objects when making adjustments on sets having an elevated mounting.

Test Equipment

The test equipment used in the servicing of television receivers should include the following:

1. Signal generator.
2. Oscilloscope.
3. Vacuum-tube voltmeter or 20,000 ohms/volt meter.
4. Tube tester.
5. Hand tools, soldering irons, etc.

The signal generator should have a frequency range of 8 to 220 mc with provision for a.m. and f.m. and separate audio output. This generator can be the one used as the marker generator for alignment adjustments.

The oscilloscope can be any conventional type used for test purposes. However, for observing waveforms an oscilloscope with a wide-band frequency response will give a more accurate indication.

The vacuum-tube voltmeter should have a low-scale range in the order of 3 volts and an input impedance of at least 10 megohms. External multipliers should be available when measuring higher voltages beyond the range of the VTVM. It will also be helpful if the VTVM has an r-f probe which may be either of the diode type or the crystal type. High-voltage test leads should be used in making voltage measurements in high-voltage circuits.

Any good tube tester with facilities for testing all the later-type tubes will be suitable. It should be kept in mind, however, that checking a tube in a tube tester is not always conclusive. In many instances, tubes such as the 6AC7, 6AB7, 6J5, and 6N7 which are used in television receivers have tested satisfactorily in tube testers of the better grade, but performed very unsatisfactorily when placed in some circuits of the receiver. However, these same tubes when placed in other circuits gave the desired performance. For example, a 6N7 tube may not operate properly as a sync separator, but will be all right as a deflection oscillator and discharge tube; or a tube that will not operate satisfactorily as an oscillator will function in an amplifier stage. Therefore, the final test of tube merit will always be its actual performance in the circuit.

A word of caution is in order here. Before inserting a new tube into a suspect stage, be sure that conditions within the circuit will not burn out the new tube. If the tube that has been removed from the set is not burned out, it can be assumed that insertion of the new tube will not damage it, and the substitution test of tube merit can be made in satety.

Solder connections in television receivers are very critical and must be made carefully. Do not use an iron so hot that the solder crystalizes, nor so cool that resin joints result. It is recommended that the hot iron be used to heat the joint and that the solder be applied directly to the heated work so that it flows evenly into the connection.

Localizing the Trouble

When a serviceman receives a call to repair a television set, he usually receives some hint as to the *possible* location of the defect. He may be informed that no picture and/or sound can be obtained, or that the quality of one or both is poor. Though trouble indications given by the set owner cannot be expected to be very accurate and cannot be relied upon to any great extent, the owner will be aware that the sound and picture are not coming through satisfactorily. Since it is quite possible that the trouble may be originating at the transmitter, or that interference, such as diathermy, may be distorting the picture temporarily, the serviceman should have available in the shop a television receiver that can be used for monitoring and should check whether or not the trouble is coming in from the transmitter of the station being received.

If examination of the set reveals that the defect is caused by conditions external to the receiver, such as interfering signals from other sources, repositioning of the antenna or transmission line may be required. In the case of diathermy very little can be accomplished other than an explanation to the set owner regarding the cause of the defective condition. In any event the serviceman must determine definitely whether or not the defect is inside the television receiver and proceed accordingly.

After the defect has been localized to the receiver proper, the next step is to determine which of the basic five sections within the receiver is not operating normally. For facility in isolating the defective stage, the receiver can be divided into sections consisting of the front end, sound, video, sweep, and power supply sections, as shown in the block diagram of Fig. 12-1. Quite often the defect can be isolated to one of the sections by observation of the sound and picture and without removing the chassis from the cabinet.

If the sound comes through clearly but no picture appears on the picture tube, this indicates that the video, sweep, or high-voltage section is defective. The front end which supplies the signal to the sound section, as well as to the video section, is evidently operating normally; this is also true of the low-voltage power supply which supplies operating potentials to all these sections. If a raster can be obtained on the picture tube by turning up the brightness control, the trouble is quite definitely isolated to the video section. If no raster is obtained, then the trouble very likely is in either the sweep or the high-voltage sections. Proper high-voltage supply will be denoted by some sort of indication on the picture tube, whereas lack of high voltage will result in no indication.

If no sound is present, or if the sound is distorted, but a normal picture is received, then the trouble is isolated to the sound section, inasmuch as normal reception of the picture indicates proper operation of the other sections of the receiver. However, defects in the sound signal can occur in the front end without noticeably affecting the picture. An example of this would be a slight detuning of the oscillator.

If no sound or picture is received, the trouble must be in some section of the receiver that is common to both. This can be either the front end or the low-voltage power supply sections. If turning up the brightness control provides a raster on the picture tube, the low-voltage supply is eliminated as the source of trouble which is then isolated to the front end. Of course, in sets incorporating a separate low-voltage supply for the sweep section, this test is valueless.

Lack of both sound and picture combined with no raster on the picture tube indicates a completely dead set, and this is probably caused by a blown fuse, defective power transformer, defective line switch or cord, or no line voltage.

With experience, the serviceman will develop a keener sense for interpreting the indications observed in defective television receivers and the ability for speedy localization of the defective section simply by looking at the picture, listening to the sound, and manipulating the panel controls.

When the serviceman has isolated the trouble to a particular section of the receiver, his next step should be to check the tubes in the defective section, inasmuch as the tubes are the first suspects in any inoperative receiver. Often this can be done without removing the chassis from the cabinet by reaching in and feeling for a cold tube. On occasion, a whole series of tubes may be found cold indicating series wiring of their heaters; so that the whole series has gone dead as a result of one tube's burning out. If no burned-out tube is found, all the tubes in the suspected section should be tested.

If the trouble cannot be determined by an observation of the picture and sound, it is necessary to remove the chassis and perform a series of systematic checks with test instruments to isolate the defective section.

These systematic checks are performed at significant points in the receiver by means of voltage or resistance measurements, or by injecting a signal into the input of a section and observing the output indication. The output indication can be observed either on the picture tube, the test oscilloscope, or by listening to the speaker of the receiver. The checks should be performed in the indicated sequence, eliminating one section at a time until the defective section is discovered. Once a defect has been localized to a particular section, further checks using signal-tracing methods must be performed to determine the faulty stage within the section. After the trouble has been isolated to a particular stage, conventional voltage and resistance measurements should be made to determine the defective component.

The systematic procedure for isolating the defective section is as follows:

Check 1. Low-Voltage Power Supply. Using the voltmeter, check for proper voltage at the output of the power supply.

Check 2. High-Voltage Power Supply. Using the VTVM and external multipliers, if necessary, check for proper voltage at the output of the high-voltage power supply. *Caution:* Use extreme care when making this check.

Check 3. Sweep Section. Connect the oscilloscope to the horizontal and vertical deflecting plates, in turn, (or to the deflecting yoke terminals) and check for proper waveform and peak-to-peak voltage. These should conform to the reference waveform and peak voltages for the particular receiver.

Check 4. Video Section. Set the signal generator to the video i.f. and connect it to the input of the video i-f section. Turn on the a.m. of the signal generator and check for a series of light and dark bars in the picture tube.

Check 5. Sound Section. Set the signal generator to the frequency of the sound i-f section and connect it to the input of the sound i-f section. Set for f.m. with a deviation of ±25 kc and listen for a tone in the speaker.

Check 6. Front End. Set the signal generator to the sound carrier frequency of the channel to which the set is tuned and connect it to the antenna input terminals. Set for f.m. with a deviation of ±25 kc and listen for a tone in the speaker. An alternative method is to set the signal generator to the video carrier frequency with a.m. and check for light and dark bars in the picture. If necessary, either of these two check methods may be repeated for each of the other channels.

Check 7. Sync Circuits. A further check may be required in the sync circuits to determine whether or not these circuits are operating properly. Connect the test oscilloscope to the output of the sync separator, or the sync amplifier tube if one is used. Check for proper waveform and peak-to-peak voltage of the sync signals. These should conform to the reference waveform and peak voltages for the particular receiver.

Correct voltages and waveforms at the stated check points will vary with the type of television receiver. Obviously, it is not possible in a general servicing procedure such as this to include reference waveforms, voltages, and resistance measurements for the many types of television receivers. In some cases, it may be possible to service a defective television receiver without these reference data. However, if this information is available, it will facilitate servicing and greatly reduce the amount of time spent in repairing a receiver. These reference waveforms, voltages, and resistance measurements may be found in the complete data on current television receivers contained in Rider's Television Manual 1.

When it has been established that the defect is in a particular section of the television receiver it must then be traced to a specific stage. In the high- and low-voltage power supply sections, voltage and resistance measurements are made to isolate the fault. However, in the sweep, video, sound, and front-end sections, signal tracing will facilitate the location of the faulty stage.

If an abnormal indication has been obtained at the output of either the vertical or horizontal sweep circuits, the oscilloscope should be moved back progressively through the section from the output to the input of each stage until a normal indication appears on the oscilloscope. The defect will lie in the stage immediately following the point where the normal indication is obtained. Further voltage and resistance measurements will isolate the defective part within the stage. This part should be replaced with one having the same characteristics.

If the defective circuit is traced to the video section, the stage can be isolated by the following procedure. The first step is to determine whether or not the trouble lies ahead of the video detector or in the stages following it; that is, whether it is in the video i-f or in the video amplifier stages. This can be done by connecting the oscilloscope to the output of the video detector and the signal generator, amplitude modulated and set to the video i.f., to the input of the video i-f strip. A detected audio signal should be obtained on the oscilloscope if the section is operating normally up to the video detector. If an abnormal indication is obtained, the signal generator should be connected to the grid of the final video i-f amplifier. If the output is still abnormal, then a check of the final video i-f amplifier and video detector stages is in order. If a normal indication is obtained at the final stages, the signal generator should be moved back stage by stage until the normal indication is no longer obtained, which will indicate that the signal generator is connected to the defective stage While moving through the video i-f stages, a rough check on the gain per stage can be made by noting how much the signal generator output must be reduced in order to maintain the same amplitude of signal on the oscilloscope.

If a normal indication from the video detector is obtained with the signal generator connected to the input of the video i-f strip, then the fault quite obviously lies in the stages following the detector. In which case the oscilloscope should be connected to the grid of the picture tube. Set the signal generator for 400-cycle output and connect it to the grid of the video output tube. If an abnormal indication is obtained on the oscilloscope, the trouble is in the video output stage. If a normal signal is obtained, the signal generator should be moved back to the grid of each video amplifier stage in turn and the output observed. If all video amplifier stages check normal, the signal generator should be left at the grid of the first video amplifier and set for a high-level 400-cycle output. A pattern of light and dark bars should appear on the picture tube. If this pattern is absent, and if the sweep section and both power supplies check normal, a bad picture tube is indicated.

The sound i-f section can be checked similarly to the video i-f section. However, the signal generator is set to the sound i.f., and f.m. is used. The oscilloscope is connected to the output of the discriminator, and the trouble is isolated to either the sound i-f stages or to the audio amplifier stages that follow the discriminator. Localization of the defective stage can be accomplished as before. However, in checking the audio amplifier stages the loudspeaker may be used as an output indicator, unless the loudspeaker itself has been found defective.

If checks 1 to 5 of the procedure for isolating the defective section are performed with normal results, the fault would appear to be in the front end of the receiver. Check 6 of this procedure will provide definite information in this direction. If no tone can be obtained in the speaker with the signal generator connected to the antenna input terminals and set to the sound carrier frequency (frequency modulated), move the signal generator to the grid of the mixer tube. If a normal response is obtained, the trouble definitely lies in the r-f amplifier stage. If a normal response is not obtained, the fault is in the mixer or oscillator stages. A quick check of the oscillator can be performed by measuring the bias voltage developed across the grid resistor; exceptionally low or no bias voltage indicates a defect in the oscillator stage. If the oscillator checks normal, the fault is in the mixer stage, and voltage and resistance measurements, as previously stated, should be made to determine the faulty part in the circuit.

Troubleshooting Chart

The following troubleshooting chart is included to supplement the systematic servicing procedure previously outlined. When the defect has been traced to a particular stage, the chart may be used to check first those parts in the circuit which are more subject to breakdown. This chart cannot include every possible type of trouble that can be encountered in servicing television receivers, but an attempt has been made to cover those troubles which experience has shown are most likely to develop. This chart covers the receiver by sections, as in the case of the localizing procedure. These sections are the low-voltage power supply, high-voltage power supply, sweep, video, sound, and front-end sections. The chart is divided into sections: each lists the faulty indication, and below this are listed the most probable causes of the observed symptom. In a number of instances, the probable causes listed will apply only to particular types of receivers incorporating specific circuits. For example, in the high-voltage power supply reference made to an oscillator coil will apply only to an r-f type of high-voltage supply, and not to a kick-back or 60-cycle type of supply. A high-voltage transformer will be found only in a 60-cycle type of supply. The fact that no low voltage is present will not affect a 60-cycle type of power supply but will render the other two types inoperative. In utilizing this chart, the serviceman will select only those probable causes that apply to the particular receiver under test.

In making checks on suspected components, the conventional voltage and resistance measurements, supplemented with signal-tracing methods, are used to determine the defective part. It should be stressed here that this chart does not represent a cure-all, but is meant to supplement the regular servicing procedure and to afford a quick check of the most probable causes of circuit failure.

TROUBLESHOOTING CHART

1. LOW-VOLTAGE POWER SUPPLY

Faulty Indication

No Output Voltage

Probable Causes:

No line voltage
Defective line switch
Blown fuse
Defective power transformer
Defective rectifier tube
Short-circuited filter capacitor
Open choke
Loose or broken connections
Defective interlock switch

Faulty Indication

Low Output Voltage

Probable Causes:

Low line voltage
Open filter capacitor
Leaky filter capacitor
Weak rectifier tube
Filter capacitors changed value
Poor solder connections
Shorted bleeder resistor
Grounded choke

2. HIGH-VOLTAGE POWER SUPPLY

Faulty Indication

No Output Voltage

Probable Causes:

No low-voltage output
Defective tubes
Defective high-voltage power transformer
Shorted filter capacitor
Open filter resistor
Open oscillator coil
Open sweep output transformer
No horizontal sweep voltage (check sweep circuit)
Loose or broken connection

Faulty Indication

Low Output Voltage

Probable Causes:

Defective low-voltage supply
Weak rectifier tube
Weak oscillator tube
Leaky filter capacitor
Open filter capacitor
Leaky sweep output coupling capacitor
Poor solder connection
Insulation breakdown

3. SWEEP SECTION

Faulty Indication

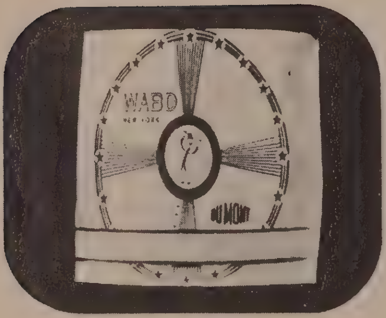
No Horizontal or Vertical Sweep

Probable Causes:

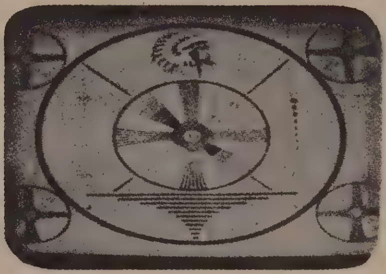
Defective tubes
Open coupling capacitors
Open or shorted bypass capacitors
Open blocking-oscillator transformer
Open resistors
Open deflection coils
Open sweep output transformer
Loose or broken connections

Faulty Indication

Insufficient Sweep Amplitude



Horizontal Width Control Misadjusted.



Vertical Height Control Misadjusted.

Probable Causes:

Height or width control improperly adjusted
Weak tubes
Resistors have changed value
Defective deflection coils
Defective sweep output transformer

Faulty Indication

No Raster — Sound Satisfactory

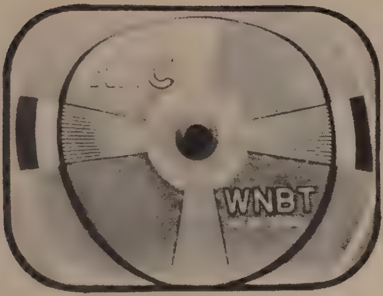
Probable Causes:

No high voltage to picture tube
Defective picture tube
Beam-bending coil improperly adjusted
Beam-bending control improperly adjusted
Loose or broken connection

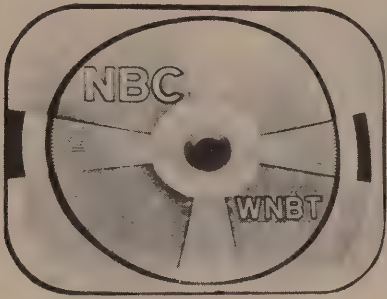
3. SWEEP SECTION — Continued

Faulty Indication

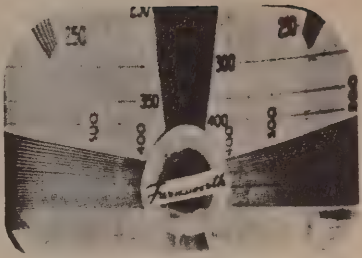
Nonlinear Horizontal or Vertical Sweep



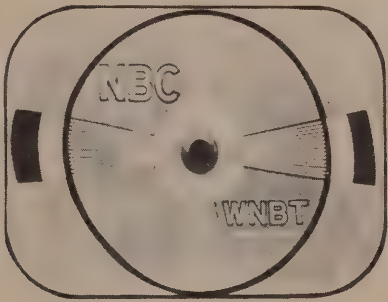
Vertical Linearity Control
Misadjusted.



Horizontal Linearity Control
Misadjusted.



Vertical Linearity Control
Misadjusted.



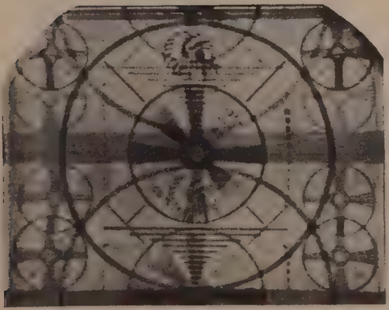
Horizontal Linearity Control
Misadjusted.

Probable Causes:

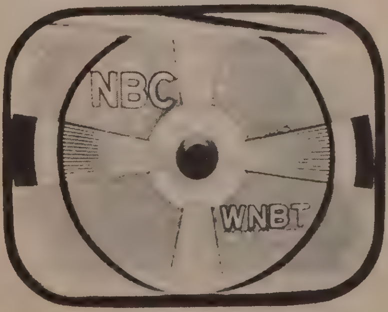
Horizontal or vertical linearity controls improperly adjusted
Defective tubes
Defective sweep output transformer
Defective deflection coils or shunt resistors
Capacitors have changed value

Faulty Indication

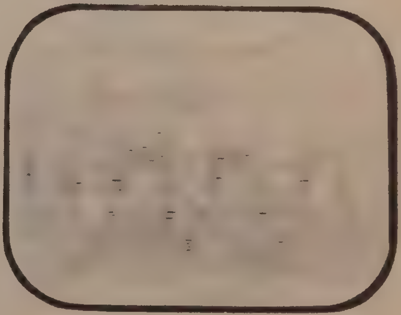
No Synchronization of Horizontal or Vertical Sweep



Vertical Hold Control
Misadjusted.

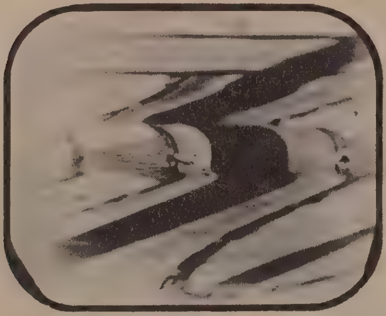


Horizontal Hold Control
Misadjusted.

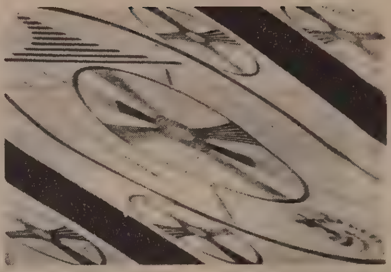


Horizontal Hold Control
Misadjusted.

3. SWEEP SECTION — Continued



Horizontal Hold Control Misadjusted.



Horizontal Hold Control Misadjusted.

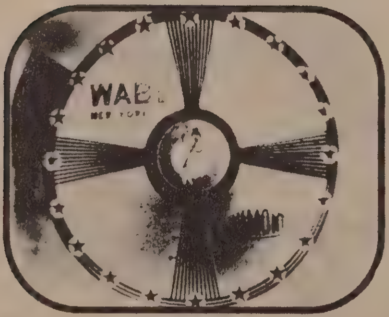
Probable Causes:

Horizontal or vertical hold controls improperly adjusted
Insufficient signal pickup at the antenna
Low gain in video i-f section
Defective sync separator stage
Defective tubes in sync circuits
Open coupling capacitor
Resistors or capacitors have changed values in sweep oscillators

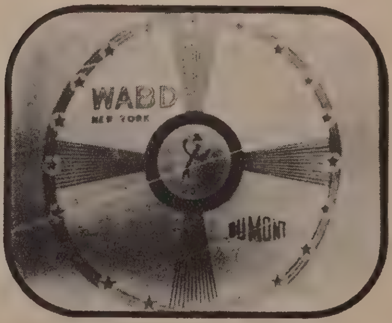
4. VIDEO SECTION

Faulty Indication

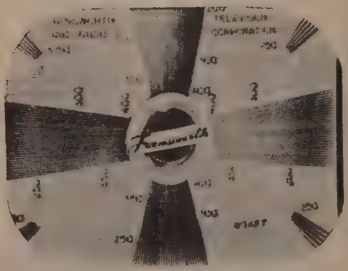
Poor Picture Definition



Contrast Too High.



Contrast Too Low.



Misalignment or Improper Antenna Orientation.

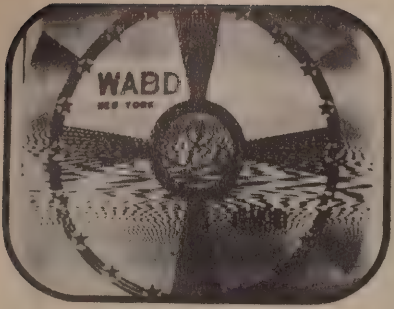
Probable Causes:

Improper adjustment of focus, contrast, or brightness controls
Weak input signal
Incorrect channel tuning adjustment
Weak tubes
Incorrect alignment of video i-f stages
Defective peaking coils

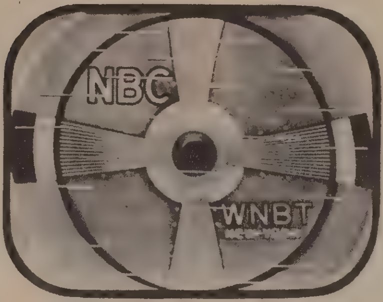
4. VIDEO SECTION — Continued

Faulty Indication

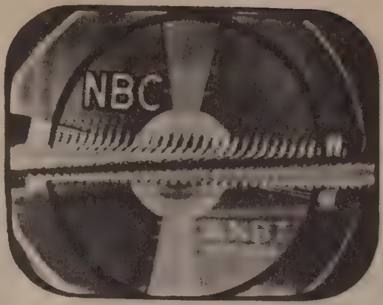
Noise in Picture



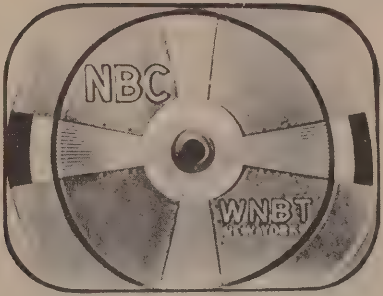
Strong Diathermy Interference.



Ignition Interference.



Diathermy Interference.



Transients.

Probable Causes:

Interference due to external causes, such as, diathermy, ignition, electric equipment, etc.
Dirty or worn contacts on switches, sockets, etc.
Noisy tubes
Noisy resistors or capacitors

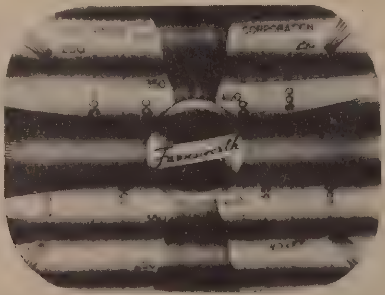
Faulty Indication

Sound in Picture

(Indicated by lines or bands in picture which vary with the audio signal.)

Probable Causes:

Incorrect trap alignment
Oscillator improperly adjusted
Oscillations in video i-f stages
Sound and video i-f leads too close together



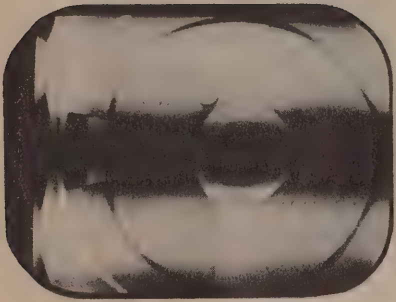
Sound Bars or Microphonics.

4. VIDEO SECTION — Continued

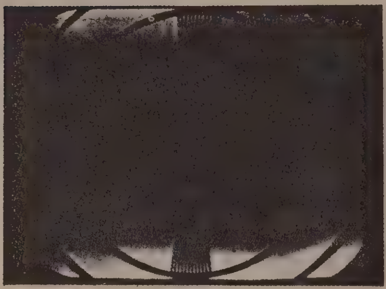
Faulty Indication

Hum in Picture

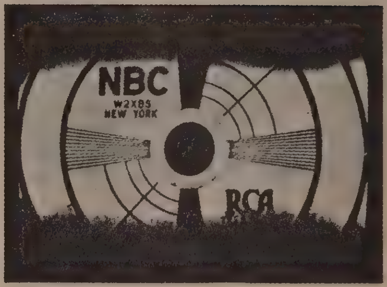
(Wide dark horizontal bands.)



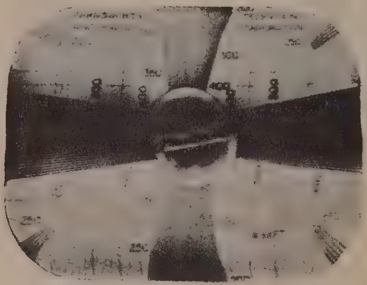
Hum in Video and Sync



Excessive Ripple in Video Amplifier.



Excessive Ripple in Video Amplifier.



120-Cycle Hum in Video and Horizontal Scanning.

Probable Causes:

Open or leaky filter capacitor in low-voltage supply
Defective tubes
Defective decoupling filters in video amplifier stages
Pickup due to misplaced leads

Faulty Indication

Smeared Picture

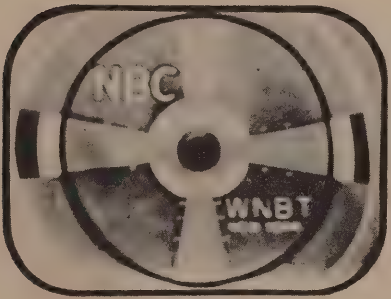
Probable Causes:

Defective load resistors or peaking coils in video amplifier stages
Open bypass capacitors
Video i-f stages misaligned
Excessive signal input to receiver

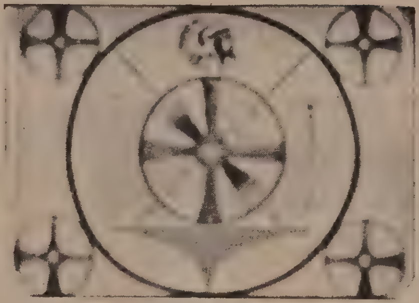
4. VIDEO SECTION — Continued

Faulty Indication

Blurred Picture



Focus Misadjusted.



Focus Misadjusted.

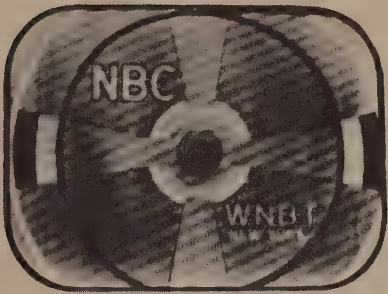
Probable Causes:

Improper adjustment of focus control
Defective focus coil

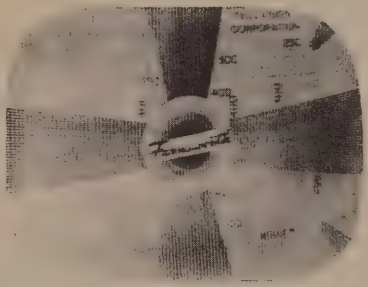
Defective focus control

Faulty Indication

Narrow Vertical or Diagonal Lines in Picture



Interference from Another Signal.

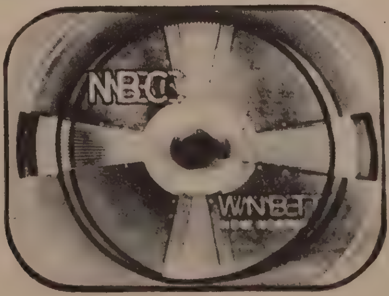


Beat Frequency.

Probable Causes: Interference from nearby radio transmitters for other services such as amateur, communications, f.m., etc.

Faulty Indication

Ghosts



Ghosts.

Probable Causes:

Poor antenna orientation
Reflector element required

5. SOUND SECTION

Faulty Indication

Weak or Distorted Audio

Probable Causes:

- Fine tuning control improperly adjusted
- Defective tube
- Open filter capacitor
- Open coupling capacitor
- Open bypass capacitor
- Sound i-f or discriminator stages misaligned
- Oscillator misaligned
- Defective speaker
- Faulty connections

Faulty Indication

Noise in Audio Output

Probable Causes:

- Dirty or worn channel-switch contacts
- Noisy tube
- Overheated resistors
- Faulty capacitors
- Defective speaker
- Arcing in high-voltage circuits
- Dirty or worn volume control
- Loose tube-socket contacts
- Faulty connections

Faulty Indication

Hum in Audio Output

Probable Causes:

- Open filter capacitor
- Defective tube
- Poor ground connections
- Pickup due to misplaced leads

6. FRONT END

Faulty Indication

No Sound and No Picture; Raster OK

(on all channels)

Probable Causes:

- Defective tubes
- Open antenna input circuit
- Defective channel switch
- Open coupling capacitors
- Short-circuited bypass capacitors
- Open load resistors
- Faulty connections

6. FRONT END — Continued

Faulty Indication

No Sound and No Picture; Raster OK

(on one channel)

Probable Causes:

Defective antenna, r-f, or oscillator coil corresponding to dead channel

Oscillator for this channel incorrectly aligned

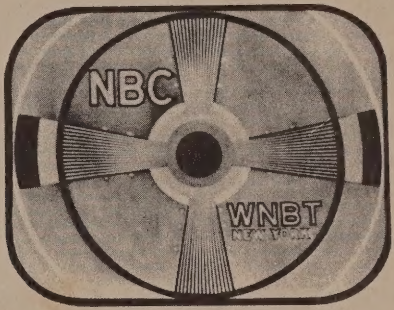
Faulty Indication

No Sound; Picture Satisfactory

(on one channel)

Probable Causes:

Oscillator for this channel slightly misaligned



Normal Picture.

The patterns used in the Troubleshooting Chart are reproduced through the courtesy of the following companies: Radio Corporation of America, General Electric Co., Allen B. Du Mont Laboratories, Inc., Farnsworth Television and Radio Corporation, Electro-Technical Industries, Motorola, Inc.